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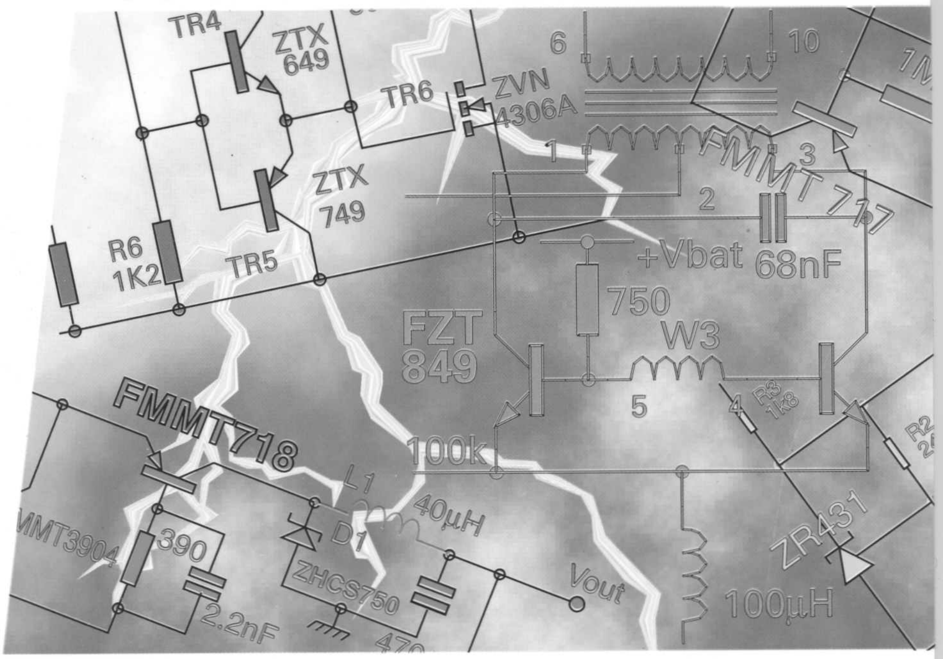
Applications Handbook



Discrete Components
Linear ICs

- Design Guidelines
- Thermal Characteristics
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ZETEX

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ZETEX

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Applications Handbook

Introduction

Welcome to the first edition of the Zetex Applications Handbook. This book contains a collection of Application and Design Notes covering a selection of the Zetex product range, including new product releases and so called "mature" products. These notes are essentially an edited compilation of existing application notes, new reports, design notes and ancillary material such as thermal data and cross references. The style of the notes ranges from discussion, through conceptual to design, and primarily is focused on the advantages afforded by the Zetex components, in terms of both performance and cost. This compilation attempts to provide a broad overview of the types of products available from Zetex, as well as examples of the application sectors and functions that benefit. We cannot claim (nor would we hope) that this treatment is comprehensive, but as the notes tend to address basic function, or fundamental circuit building blocks, it is expected that the information offered will be applicable to many sectors.

The emphasis of the notes is on practical application of the devices leading to successful and painless design, rather than on theoretical and mathematical treatments, though references are occasionally suggested for further reading. Waveforms are provided in some cases, especially where this would support circuit understanding and credibility of text assertions.

It is hoped that this handbook will prove useful in providing the most appropriate solution for a given application, thereby leading to efficient and marketable end products. To this end, and for future releases, Zetex would appreciate any feedback on its products, or the information contained within this book and supporting databooks, and is willing to consider publication of any guest application material, whether that is system, circuit or device related.

Applications Handbook

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Application Note Abstracts

AN1 ***8W Fluorescent Lamp Inverter - Compact DC to AC Inversion for Efficient Lighting Systems***

An enhanced TO92 packaged one transistor design that uses a minimum of components, and yet achieves performance and efficiency normally associated with much larger packaged switching transistors. The design is optimised to provide a fast turn-off condition for the transistor, thereby allowing the inclusion of the 80V, 2A DC rated ZTX652 yielding efficient and reliable operation. Typical waveforms and full circuit details are provided.

AN2 ***Portable Ni-Cd Battery Charger - Inexpensive Load Tolerant Step-up Energy Source***

A relatively simple low parts count topology is presented for a one transistor design, that allows "in the field" re-charge of Ni-Cd battery packs from a convenient energy source such as an automotive battery. The note includes transformer design and winding details. Though designed for a specific case, the circuit and the principle employed should lend itself to many other applications.

AN3 ***Infra Red Remote Control and Data Transmission - An Introduction to Photodiodes - Load Circuits and Applications***

A description of the optimum operating conditions for silicon p-i-n photodiodes, and examples of suitable biasing networks are provided. The note also includes a number of application circuits derived from case studies.

AN4 ***Centre Collector E-Line Applications - An Overview of Package Capability and Applications Advantages***

An introduction to the centre collector frame option of the E-Line package, including thermal discussion and argument for replacing larger packaged components.

AN5 ***Super E-Line Applications in Automotive Electronics - Replacement of Large Packaged Transistors with an Enhanced TO92 Product***

Load drivers within automotive systems have to tolerate extreme environmental and electrical conditions. This note describes how the Silicone encapsulated E-Line range can provide cost and performance benefits to the Automotive systems designer.

AN6 ***The Use of Zetex E-Line Transistors in Motor and Solenoid Driver Functions within Printers - Effective Logic to High Current Load Interfacing***

A brief discussion of the drive requirements of relatively high current inductive loads within computer peripheral equipment, with respect to the benefits possible with optimised small packaged components.

AN7 ***The Use of Zetex E-Line Transistors in DC-DC Converters - An Introduction to the E-Line Package and Typical Applications***

A basic guide to the DC-DC Converter requirement and concept, including an introduction to the Zetex enhanced TO92 - the E-Line package, and some suggested applications where these components offer advantages.

AN8 ***The ZTX415 Avalanche Mode Transistor - An Introduction to Characterisation, Performance and Applications.***

Avalanche mode switching operation of bipolar transistors is a useful but sometimes frustrating experience due to the lack of suitably designed components, and an understanding of their limiting values. The ZTX415 is the industry's first device that has been designed specifically for this function, and this note provides parametric background, typical application circuits, and also lists useful references.

AN9 ***Zetex Variable Capacitance Diodes***

An introduction to the basic theory, an explanation of the important parameters, and typical applications are described within this note. An overview is also provided for the Zetex range of variable capacitance diodes.

AN10 ***Automotive Relay Drivers Using the ZVN4206AV - Protection Free Interfacing with Avalanche Rated MOSFETs***

An alternative solution is provided to the usual large package Darlington effected relay driver. The ZVN4206AV has been designed, processed and rated to withstand significant energy within the breakdown region. This allows the part to absorb back-emf and transient energy, thereby allowing lower component count circuits.

AN11 ***Features and Applications of the FMMT618 and 619 - High Current SOT23 Replace SOT89, SOT223 and D-Pak.***

This note introduces the FMMT618 and 619 - two NPN variants of the Zetex SuperSOT family, and includes parametric overview, thermal capability discussion, brief description of the Matrix geometry, and applications.

AN12 ***The FMMT718 Range, Features and Applications - Replacing SOT89, SOT223 and D-Pak Products with High Current SOT23 Bipolar Transistors***

Following on from AN11, this note introduces other members of the Zetex SuperSOT series, and concentrates on the PNP options and complementary NPN/PNP designs.

AN13 ***The ZCN0545A and ZCP0545A Low Power IGBTs - A Silicon/Package Efficient Device for Compact Fluorescent Lamp (CFL) Ballasts and Static Switching***

These devices represent the industry's first TO92 packaged IGBTs, and have been designed for specific use in relatively high frequency switching applications, particularly in half bridge series resonant power supplies. This note introduces the characteristics, drive requirements, and includes several applications examples.

AN14 ***Transistor Considerations For LCD Backlighting - High Efficiency DC to AC Conversion.***

A study of the parametric requirements of Bipolar transistors for the self-oscillating resonant "Royer" inverter with respect to LCD backlighting, including circuit behaviour, the voltage ratings of most relevance to ensure optimised device selection, and compact packaging options.

Abstracts (Continued)

AN15 Features and Applications of the FMMT617 and FMMT717 "SuperSOT" SOT23 Bipolar Transistors - 3A NPN and 2.5A PNP SOT23 Bipolar Devices

An introduction to the low voltage variant members of the Zetex SuperSOT family, including a description of the high current capability, a comparison of the extremely low losses evident when compared with competing MOSFET options, and suggested applications.

AN16 Automotive and Household Siren Driver Circuits - ZSD100 and Discrete 'H'-Bridge Minimum Parts Count Solution

A solution is provided that uses a minimum number of components by utilising an application specific signal generation IC, and a high efficiency 'H'-bridge effected with 3A DC rated Super-β TO92 style switching transistors.

AN17 Emergency Lighting Systems and Battery Powered Fluorescent Lighting - High Current TO92 Switching Transistors Provide 87% DC-AC Conversion Efficiency

An overview of the principles behind typical emergency lighting systems, fluorescent tube behaviour, and the requirements of the transistors for efficient voltage inversion. The note also includes two "Royer" based designs for low voltage systems.

AN18 Power MOSFET Gate Driver Circuits using High Current Super-β Transistors - 6A Pulse Rated SOT23 Transistors for High Frequency MOSFET Interfacing.

A description of the MOSFET converter circuits' requirement for a high current buffer stage, a suggested range of Zetex high current capable, fast switching transistors, and circuit examples.

AN19 Component Reliability

This note introduces basic reliability theory, failure rate analysis, and product qualification methods, and includes sample data and FITs values for a variety of test methods.

AN20 Magnetoresistive Sensors - Theory of Operation and Applications

An Introduction to Magnetoresistive sensor technology as used for current and magnetic field sensor products, including theory of operation, a description of parametric terms, and application circuit examples.

AN21 Bipolar Transistor Considerations for Battery Powered Equipment - Leading to Efficiency and Competitive Advantages in Portable Systems

Bipolar products provide many performance advantages in modern battery powered systems. This note reviews the parametric and circuit related characteristics of high performance Bipolar devices and MOSFETs, and provides some application examples in which the Bipolar option gives the most credible solution.

AN22 High Frequency DC-DC Conversion using High Current Bipolar Transistors - Conversion upto 400kHz using Optimised Geometry Bipolar Devices

A discussion of the switching capabilities of Bipolar transistors, including charge analysis and circuit techniques to permit high frequency saturated switching. Circuit examples are presented to illustrate speed-up methods, and charts to demonstrate the resultant efficiency improvements. The report also includes turn-off charge versus bias point characterisation charts, for typical high current low $V_{CE(sat)}$ Bipolar devices, targeted for low voltage and battery applications.

AN23 Zetex SPICE Models - Understanding Model Parameters and Applications Limitations

This report describes the basic Gummel-Poon model as used as a basis for analog circuit simulations. An overview of the model parameters and their derivation, as well as guidelines for model manipulation for specific cases are also included. A number of typical application circuits are suggested with the resultant SPICE predictions of operation.

AN24 An Introduction to the SM-8 Package

The SM-8 is an eight leaded version of the industry standard SOT223 transistor package. Using a number of leadframe options allows the assembly of dual transistors, dual ICs and hybrid combinations. This report provides an introduction to package attributes and includes example applications that benefit.

AN25 Soldering Surface Mount Devices

PC board assembly with surface mount components requires due consideration of solder behaviour, thermal limitations of the devices concerned, and mechanical attach integrity. This note provides a brief overview of the pertinent factors, and suggests maximum temperatures and soldering profiles.

AN26 Fast Charging Batteries with Zetex High Current PNP Transistors and Benchmarq Controller ICs

High current, fast switching PNP transistors are capable of replacing large packaged P-Channel MOSFETs, when attention is paid to the base drive requirements. This enables space and performance efficient designs at much reduced cost.

AN27 ZR431 Application Note

An introduction to some of the many applications of the versatile ZR431 shunt regulator. Particular emphasis is given to the advantages apparent when employing the Zetex version of this popular part.

AN28 The ZLDO Series of Low Dropout Voltage Regulators

The ZLDO series of regulators provide the lowest dropout voltage currently available in their class. Application examples covering dropout and other important features of the series are discussed.

Subject Index

This index includes references to Application Notes AN1 to AN28, and Design Notes DN1 to DN40. The entries are arranged to include application sector, application circuit, device characteristic, and for most cases device name. The subjects are referenced by the appropriate note, followed by hyphenated page number/s.

Eg. For a reference to Application Note 14 page 3, this would read as AN14-3, for Application Note 3, pages 3 and 7, this would read as AN3-3,7. Many of the circuits suggest more than one suitable device, dependent on the customer specifics, and indeed this treatment is far from comprehensive. It is suggested that this book is used in conjunction with the device data books to ensure a complete view of available options.

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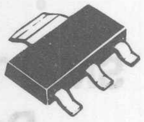
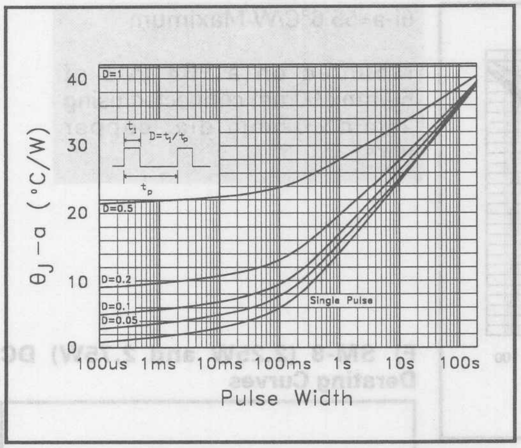
Notes

SOT223 Thermal Resistance and Derating Information



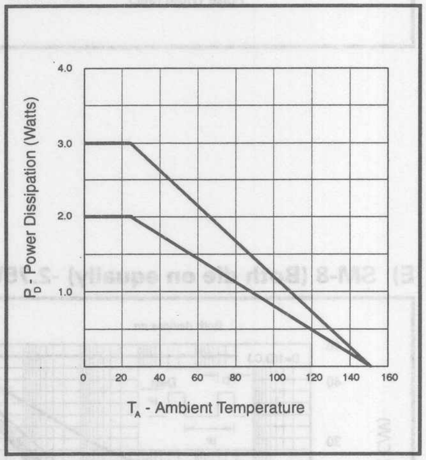
A) SOT223 3W devices



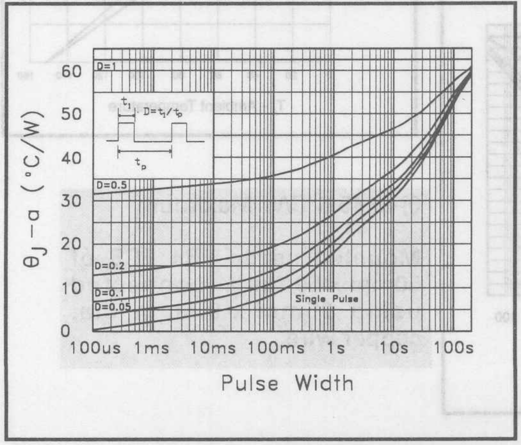
θj-a=41.7°C/W Maximum
θj-c=12°C/W Typical

Mounted on an FR4 p.c.b. of 50mm x 50mm(2" x 2") connected using 25mm x 0.5mm dia. copper wire.

C) SOT223 (2W and 3W) DC Derating Curves

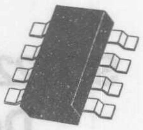


B) SOT223 2W devices



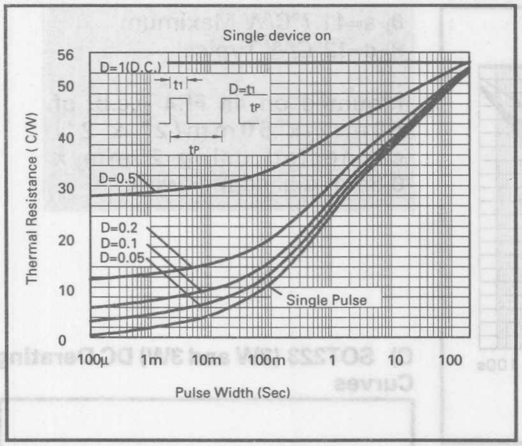
θj-a=62.5°C/W Maximum
θj-c=15°C/W Typical

Mounted on an FR4 p.c.b. of 25mm x 25mm (1" x 1") connected using 25mm x 0.5mm dia. copper wire.

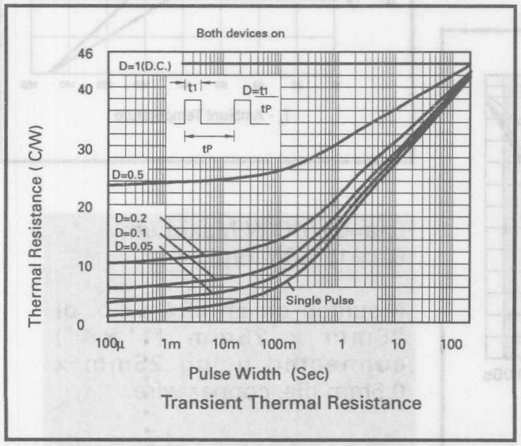


SM-8 Thermal Resistance and Derating Information

D) SM-8 (Single Die On) -2.25 W



E) SM-8 (Both die on equally) -2.75W

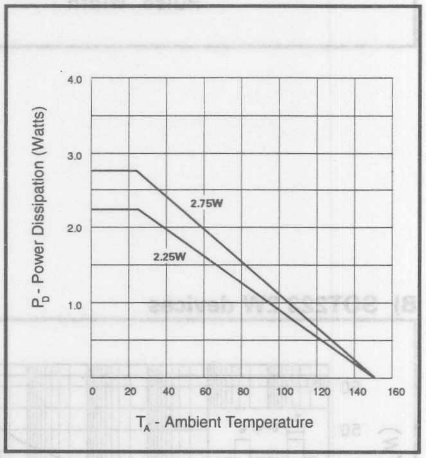


Transient Thermal Resistance

θj-a=55.6°C/W Maximum

Mounted on a FR4 PCB of 50mmx 50mm connected using 25mm x 0.5mm dia. copper wire.

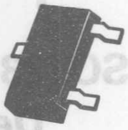
F) SM-8 (2.25W and 2.75W) DC Derating Curves



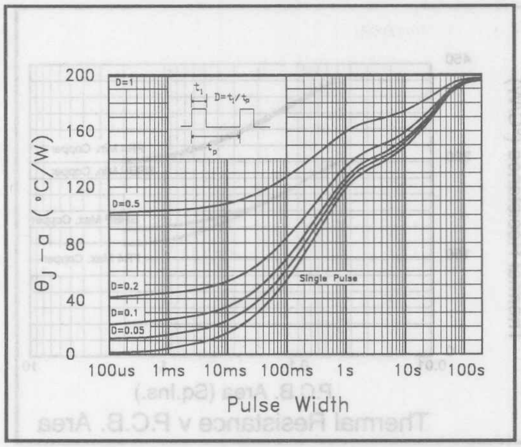
θj-a=45.6°C/W Maximum

Mounted on an FR4 PCB of 50mm x 50mm connected using 25mm x 0.5mm dia. copper wire.

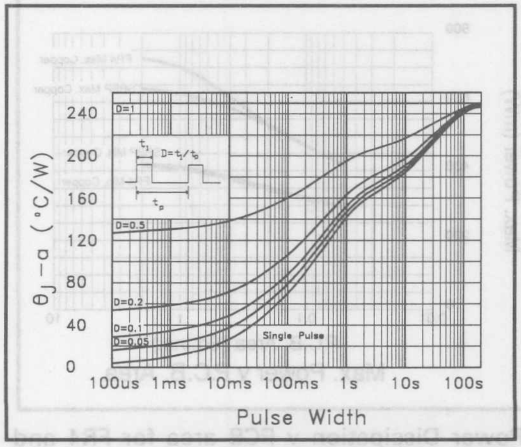
SOT23 Thermal Resistance and Derating Information



G) SOT23 625mW devices



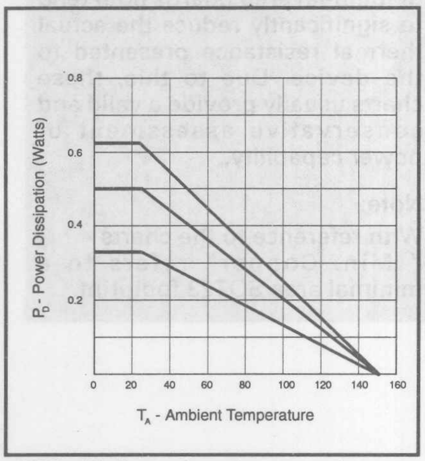
H) SOT23 500mW devices



θj-a=190°C/W Typical
=200°C/W Maximum
θj-c=100°C/W Typical

Mounted on a 15 x 15 x 0.6 mm alumina substrate connected using 25mm x 0.5mm dia copper wire.

I) SOT23 (625mW and 500mW) DC Derating Curves



*330mW shown for reference only

θj-a=200°C/W Typical
=250°C/W Maximum
θj-c=110°C/W Typical

Mounted on a 15 x 15 x 0.6 mm alumina substrate connected using 25mm x 0.5mm dia copper wire.

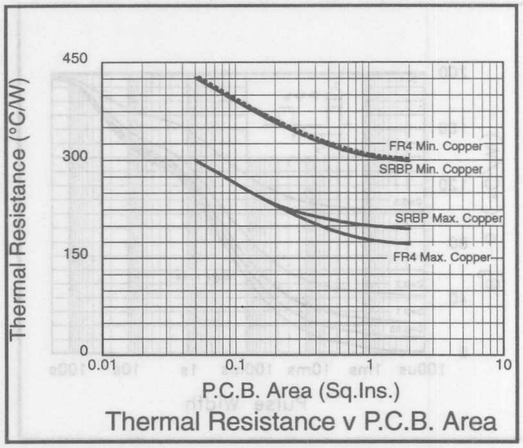
SOT23 Thermal Resistance and Derating Information

These charts illustrate how the board material, available board area and amount of copper present influence the thermal resistance experienced by a transistor.

Both charts are derived from the same measurement data, and enable an assessment of maximum junction temperature with respect to mounting method. In practical circumstances, adjacent components and the use of multi-layered boards both tend to significantly reduce the actual thermal resistance presented to the device. Due to this, these charts usually provide a valid and conservative assessment of power capability.

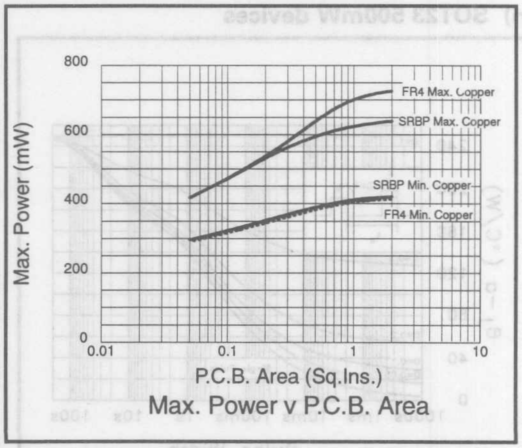
Note:
With reference to the charts - " Min. Copper" refers to a minimal area SOT23 footprint.

J) SOT23 SuperSOT (625mW) series



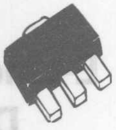
Thermal resistance (j-a) v PCB area for FR4 and SRBP (paper phenol) board material.

K) SOT23 SuperSOT (625mW) series

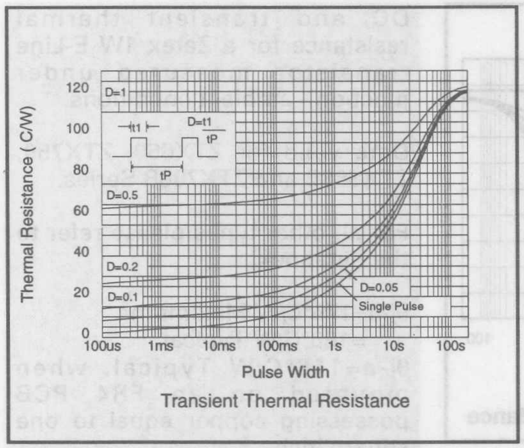


Power Dissipation v PCB area for FR4 and SRBP (paper phenol) board material.

SOT89 Thermal Resistance and Derating Information



L) SOT89 1W devices

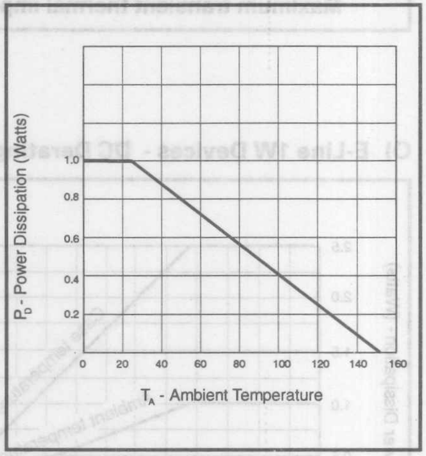


Transient Thermal Resistance

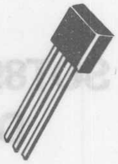
θj-a=125°C/W Maximum
θj-tab=10°C/W Typical

Mounted on a 15 x 15 x 0.6 mm alumina ceramic substrate connected using 25mm x 0.5mm dia. copper wire.

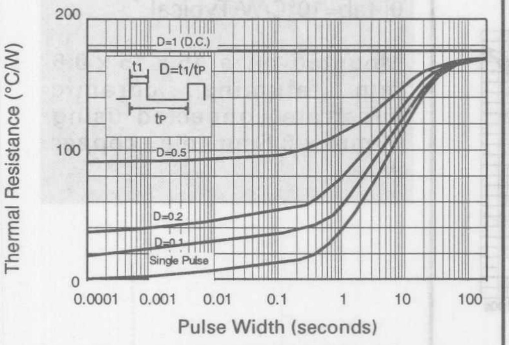
M) SOT89 (1W) DC Derating Curve



E-Line Thermal Resistance and Derating Information



N) E-Line 1W Devices



Maximum transient thermal impedance

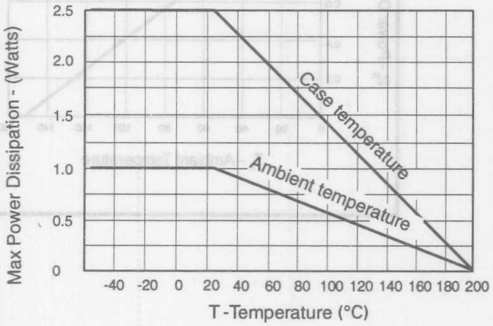
DC and transient thermal resistance for a Zetex 1W E-Line transistor measured under junction - ambient conditions.

Data valid for ZTX650, ZTX750, ZTX688B and ZTX788B Series.

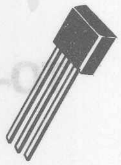
For all other types please refer to the databook.

θj-a=175°C/W Maximum
=160°C/W Typical
θj-a=116°C/W Typical, when mounted on an FR4 PCB possessing copper equal to one square inch.
θj-c=70°C/W Maximum

O) E-Line 1W Devices - DC Derating Curves

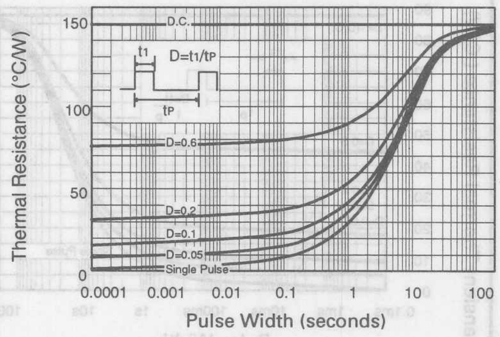


Derating curve



E-Line Thermal Resistance and Derating Information

P) E-Line 1.2W Devices



Maximum transient thermal impedance

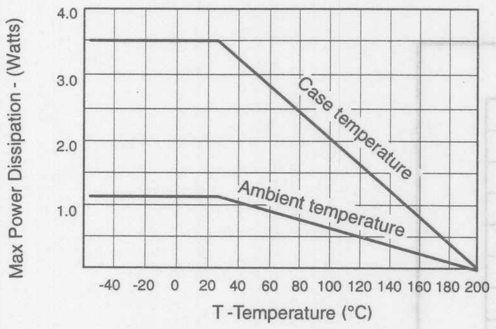
DC and transient thermal resistance for a Zetex 1.2W E-Line transistor measured under junction - ambient conditions.

Data valid for ZTX850 and ZTX950 Series.

For all other types please refer to the databook.

θ_{j-a}=150°C/W Maximum
=143°C/W Typical
θ_{j-a}=116°C/W Typical
θ_{j-c}= 50°C/W Maximum

Q) E-Line 1.2W Devices - DC Derating Curves



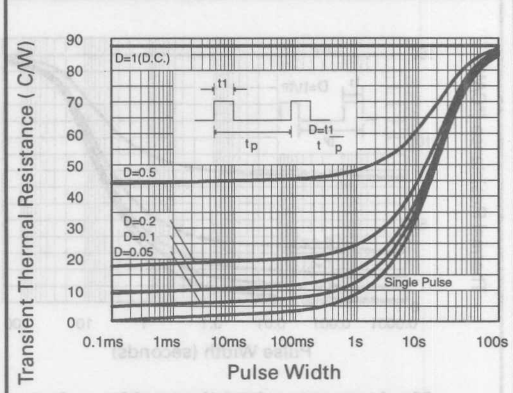
Derating curve

TO-126 Thermal Resistance and Derating Information

DC and transient thermal resistance for a Zetex 2W TO-126 transistor measured under junction - ambient conditions. Data valid for all ZBD types eg ZBD949 etc.

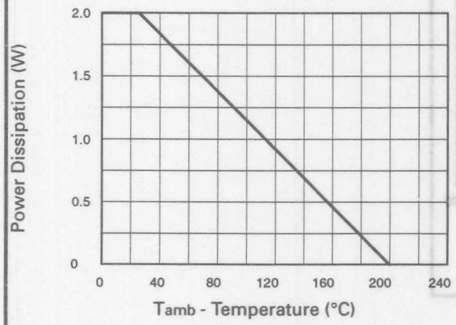
θj-a=87.5°C/W Maximum
θj-c=7°C/W Typical

R) TO-126 2W Devices



Transient Thermal Resistance (Junction To Ambient)

S) TO-126 DC Derating Curve



Derating Curve

Surface Mount Alternatives

The following tables have been generated to assist in the conversion to surface mount packages. Note that the packages offered do not necessarily possess the same power dissipation.

Please note that this does not represent a complete product listing - refer to Semiconductor Databooks: "Book 1 - Through Hole Components"; Book 2 - "High Performance Surface Mount Devices" and "High Performance Bipolar Integrated Circuits". (Details of these and other ZETEX publications can be found on pages 40 and 41).

Part No.	Surface Mount Part Number	Package
2N6517	FMMT6517	SOT23
2N6520	FMMT6520	SOT23
2N6714	FMMT449	SOT23
2N6715	FMMT491	SOT23
2N6716	FMMT491	SOT23
2N6717	FMMT493	SOT23
2N6718	FMMT493	SOT23
2N6724	FMMT38C	SOT23
2N6725	FMMT38C	SOT23
2N6726	FMMT589	SOT23
2N6727	FMMT591	SOT23
2N6728	FMMT591	SOT23
2N6729	FMMT593	SOT23
2N6730	FMMT593	SOT23
2N6731	FMMT493	SOT23
2N6732	FMMT593	SOT23
2N7000	2N7002	SOT23
2N7000P	2N7002	SOT23
BC368	FMMT449	SOT23
BC369	FMMT549	SOT23
BC372	FMMT614	SOT23
BC372P	FMMT614	SOT23
BC639	FMMT493	SOT23
BC640	FMMT593	SOT23
BCX38A	FMMT38A	SOT23
BCX38B	FMMT38B	SOT23
BCX38C	FMMT38C	SOT23
BS107P	ZVNL120G	SOT223
BS107PT	ZVNL120G	SOT223
BS170P	BS170F	SOT23
BS250P	BS250F	SOT23
FXT38C	FMMT38C	SOT23
FXT449	FMMT449	SOT23
FXT450	FMMT491	SOT23
FXT451	FMMT491	SOT23
FXT453	FMMT493	SOT23
FXT455	FMMT495	SOT23
FXT458	FMMT458	SOT23
FXT549	FMMT549	SOT23
FXT550	FMMT591	SOT23
FXT551	FMMT591	SOT23
FXT553	FMMT593	SOT23
FXT555	FMMT596	SOT23
FXT557	FMMT597	SOT23
FXT601B	FZT601B	SOT223
FXT603	FZT603	SOT223
FXT605	FZT605	SOT223
FXT614	FMMT614	SOT23
FXT649	FZT649	SOT223
FXT651	FZT651	SOT223
FXT653	FZT653	SOT223
FXT655	FZT655	SOT223
FXT657	FZT657	SOT223
FXT688B	FZT688B	SOT223
FXT690B	FZT690B	SOT223
FXT704	FZT704	SOT223
FXT705	FZT705	SOT223
FXT749	FZT749	SOT223
FXT751	FZT751	SOT223
FXT753	FZT753	SOT223
FXT755	FZT755	SOT223
FXT757	FZT757	SOT223

Surface Mount Alternatives

Part No.	Surface Mount Part Number	Package	Part No.	Surface Mount Part Number	Package
FXT788B	FZT788B	SOT223	ZC2800	ZC2800E	SOT23
FXT790A	FZT790A	SOT223	ZC2811	ZC2811E	SOT23
FXTA42	FMMTA42	SOT23	ZC5800	ZC5800E	SOT23
FXTA92	FMMTA92	SOT23	ZCN0545A	REFER	
MPS2222A	FMMT2222A	SOT23	ZDX1F	BAV74	SOT23
MPS2369A	FMMT2369A	SOT23	ZDX1R	BAW56	SOT23
MPS2907A	FMMT2907A	SOT23	ZDX2F	BAV74	SOT23
MPS5179	FMMT5179	SOT23	ZDX2R	BAW56	SOT23
MPS6601	FMMT449	SOT23	ZDX3F	FLLD258	SOT23
MPS6651	FMMT549	SOT23	ZDX3R	FLLD263	SOT23
MPSA06	FMMTA06	SOT23	ZDX4F	FLLD258	SOT23
MPSA12P	FMMTA12	SOT23	ZDX4R	FLLD263	SOT23
MPSA42	FMMTA42	SOT23	ZDX5	BAV99	SOT23
MPSA44	FZT658	SOT223	ZDX6	BAV99	SOT23
MPSA56	FMMTA56	SOT23	ZR431C	ZR431G	SOT223
MPSA77	BCV46	SOT23	ZRA124A	ZRA124F	SOT23.
MPSA77P	BCV46	SOT23	ZRA124Y	ZRA124F	SOT23.
MPSA92	FMMTA92	SOT23	ZRA125A	ZRA125F	SOT23.
MPSA94	FMMT558	SOT23	ZRA125Y	ZRA125F	SOT23.
MPSH10	FMMTH10	SOT23	ZRA245A	ZRA245F	SOT23.
MPSH10P	FMMTH10	SOT23	ZRA245Y	ZRA245F	SOT23.
VN10LP	VN10LF	SOT23	ZRA250A	ZRA250F	SOT23.
VN2222LL	ZVN3306F	SOT23	ZRA250Y	ZRA250F	SOT23.
ZBD849	FZT849	SOT223	ZRA400A	ZRA400F	SOT23.
ZBD853	FZT853	SOT223	ZRA400Y	ZRA400F	SOT23.
ZBD857	FZT857	SOT223	ZRB500A	ZRB500F	SOT23.
ZBD869	FZT869	SOT223	ZRB500Y	ZRB500F	SOT23.
ZBD949	FZT949	SOT223	ZRC330A	ZRC330F	SOT23.
ZBD953	FZT953	SOT223	ZRC330Y	ZRC330F	SOT23.
ZBD957	FZT957	SOT223	ZSR330C	ZSR330G	SOT223
ZC744	FMMV2105	SOT23	ZSR500C	ZSR500G	SOT223
ZC820/A/B	ZC830/A/B	SOT23	ZSR600C	ZSR600G	SOT223
ZC821/A/B	ZC831/A/B	SOT23	ZSR800C	ZSR800G	SOT223
ZC822/A/B	ZC832/A/B	SOT23	ZSR1000C	ZSR1000G	SOT223
ZC823/A/B	ZC833/A/B	SOT23	ZTX214C	BC860C	SOT23.
ZC824/A/B	ZC834/A/B	SOT23	ZTX300	BCW60B	SOT23.
ZC825/A/B	ZC835/A/B	SOT23	ZTX314	FMMT2369A	SOT23.
ZC826/A/B	ZC836/A/B	SOT23	ZTX320	FMMT918	SOT23.

Surface Mount Alternatives

Part No.	Surface Mount Part Number	Package
ZTX321	FMMT918	SOT23.
ZTX322	BFQ31	SOT23.
ZTX323	BFQ31A	SOT23.
ZTX325	BFS17	SOT23.
ZTX337C	BCW65H	SOT23.
ZTX341	BCX41	SOT23.
ZTX360	REFER	
ZTX384C	BC859C	SOT23
ZTX413	FMMT413	SOT23
ZTX415	FMMT415	SOT23
ZTX449	FMMT449	SOT23
ZTX450	FMMT491	SOT23
ZTX451	FMMT491	SOT23
ZTX452	FMMT493	SOT23
ZTX453	FMMT493	SOT23
ZTX454	FMMT495	SOT23
ZTX455	FMMT495	SOT23
ZTX457	FMMT497	SOT23
ZTX458	FMMT458	SOT23
ZTX500	BCW61B	SOT23
ZTX510	BSS65	SOT23
ZTX537C	BCW68H	SOT23
ZTX541	HT3	SOT23
ZTX549	FMMT549	SOT23
ZTX549A	FMMT549A	SOT23
ZTX550	FMMT591	SOT23
ZTX551	FMMT591	SOT23
ZTX552	FMMT593	SOT23
ZTX553	FMMT593	SOT23
ZTX554	FMMT593	SOT23
ZTX555	FMMT596	SOT23
ZTX556	FMMT596	SOT23
ZTX557	FMMT597	SOT23
ZTX558	FMMT558	SOT23
ZTX576	FMMT596	SOT23
ZTX600	FZT601B	SOT223
ZTX600A	FZT601B	SOT223
ZTX600B	FZT601B	SOT223
ZTX601	FZT601B	SOT223

Part No.	Surface Mount Part Number	Package
ZTX601A	FZT601B	SOT223
ZTX601B	FZT601B	SOT223
ZTX602	FZT603	SOT223
ZTX603	FZT603	SOT223
ZTX604	FZT604	SOT223
ZTX605	FZT605	SOT223
ZTX614	FMMT614	SOT23
ZTX618	FMMT618	SOT23
ZTX649	FZT649	SOT223
ZTX650	FZT651	SOT223
ZTX651	FZT651	SOT223
ZTX652	FZT653	SOT223
ZTX653	FZT653	SOT223
ZTX654	FZT655	SOT223
ZTX655	FZT655	SOT223
ZTX656	FZT657	SOT223
ZTX657	FZT657	SOT223
ZTX658	FZT658	SOT223
ZTX688B	FZT688B	SOT223
ZTX689B	FZT689B	SOT223
ZTX690B	FZT690B	SOT223
ZTX692B	FZT692B	SOT223
ZTX694B	FZT694B	SOT223
ZTX696B	FZT696B	SOT223
ZTX704	FZT705	SOT223
ZTX705	FZT705	SOT223
ZTX712	BCV46	SOT223
ZTX718	FMMT718	SOT23
ZTX749	FZT749	SOT223
ZTX750	FZT751	SOT223
ZTX751	FZT751	SOT223
ZTX752	FZT753	SOT223
ZTX753	FZT753	SOT223
ZTX754	FZT755	SOT223
ZTX755	FZT755	SOT223
ZTX756	FZT757	SOT223
ZTX757	FZT757	SOT223
ZTX758	FZT758	SOT223
ZTX776	REFER	

Surface Mount Alternatives

Part No.	Surface Mount Part Number	Package
ZTX788A	REFER	
ZTX788B	FZT788B	SOT223
ZTX789A	FZT789A	SOT223
ZTX790A	FZT790A	SOT223
ZTX792A	FZT792A	SOT223
ZTX795A	FZT795A	SOT223
ZTX796A	FZT796A	SOT223
ZTX849	FZT849	SOT223
ZTX851	FZT851	SOT223
ZTX853	FZT853	SOT223
ZTX855	FZT855	SOT223
ZTX857	FZT857	SOT223
ZTX869	FZT869	SOT223
ZTX948	FZT948	SOT223
ZTX949	FZT949	SOT223
ZTX951	FZT951	SOT223
ZTX953	FZT953	SOT223
ZTX955	FZT955	SOT223
ZTX956	FZT956	SOT223
ZTX957	FZT957	SOT223
ZTX958	FZT958	SOT223
ZTX968	FZT968	SOT223
ZTX1047A	REFER	
ZTX1048A	REFER	
ZTX1049A	REFER	
ZTX1051A	REFER	
ZTX1053A	REFER	
ZTX1055A	REFER	
ZTX1056A	REFER	
ZVN0120A	ZVNL120G	SOT223
ZVN0124A	ZVNL120G	SOT223
ZVN0535A	REFER	
ZVN0540A	ZVN0545G	SOT223
ZVN0545A	ZVN0545G	SOT223
ZVN1409A	REFER	
ZVN2106A	ZVN2106G	SOT223
ZVN2110A	ZVN2110G	SOT223
ZVN2110C	ZVN2110G	SOT223

Part No.	Surface Mount Part Number	Package
ZVN2120A	ZVNL120G	SOT223
ZVN2120C	ZVNL120G	SOT223
ZVN2535A	ZVNL120G	SOT223
ZVN3306A	ZVN3306F	SOT23
ZVN3310A	ZVN3310F	SOT23
ZVN4206A	ZVN4206G	SOT223
ZVN4206C	ZVN4206G	SOT223
ZVN4210A	ZVN4210G	SOT223
ZVN4306A	ZVN4306G	SOT223
ZVN4310A	ZVN4310G	SOT223
ZVN4424A	ZVN4424G	SOT223
ZVN4424C	ZVN4424G	SOT223
ZVNL110A	ZVNL110G	SOT223
ZVNL120A	ZVNL120G	SOT223
ZVNL120C	ZVNL120G	SOT223
ZVNL535A	REFER	
ZVP0120A	ZVP2120G	SOT223
ZVP0535A	REFER	
ZVP0540A	REFER	
ZVP0545A	REFER	
ZVP1320A	ZVP1320F	SOT23
ZVP2106A	ZVP2106G	SOT223
ZVP2106C	ZVP2106G	SOT223
ZVP2110A	ZVP2110G	SOT223
ZVP2110C	ZVP2110G	SOT223
ZVP2120A	ZVP2120G	SOT223
ZVP2120C	ZVP2120G	SOT223
ZVP3306A	ZVP3306F	SOT23
ZVP3310A	ZVP3310F	SOT23
ZVP4105A	BSS84	SOT23
ZVP4424A	ZVP4424G	SOT223
ZVP4424C	ZVP4424G	SOT223

Glossary

An explanation of the symbols used in tables and data within these Application and Design Notes. These represent measurement and circuit parameters relevant to Bipolar, MOS and diode technologies.

Symbol	**Parameter**
BV_{DSS}	Drain-source breakdown voltage
C_{CB}	Collector-base capacitance (emitter AC grounded)
C_D	Diode capacitance
C_{EB}	Emitter-base capacitance (collector AC grounded)
C_{ibo}	Input capacitance (base-emitter, collector open)
C_{iss}	Input capacitance
C_{oss}	Common source output capacitance
C_{obo}	Output capacitance (collector-base, emitter open)
$-C_{re}$	Feedback capacitance
C_{rss}	Reverse transfer capacitance
C_T	Diode reverse capacitance
C_{TC}	Collector-base capacitance
C_{TE}	Emitter-base capacitance
E_{AR}	Avalanche energy (repetitive)
ESR	Effective series resistance
f	Frequency
F_{op}	Operating frequency
f_T	Transition frequency
g_{fs}	Forward transconductance
Gpe	Common-emitter amplifier power gain
h_{fe}	Small signal current gain
h_{FE}	Static forward current transfer ratio
I_{AR}	Avalanche current (repetitive)
I_B	Base Current
I_{BEX}	Base cut-off current
I_{B1}	Control Current (Switching)
I_{B2}	Turn-off base current (Switching)
I_C	Collector current
I_{CBO}	Collector-base cut-off current (emitter open)
I_{CEO}	Collector-emitter cut-off current (base open)
I_{CER}	Collector-emitter cut-off current (with external base-emitter resistor)

Glossary

Symbol	Parameter
I_{CES}	Collector-emitter cut-off current (base-emitter shorted)
I_{CEX} or I_{CEV}	Collector emitter cut-off current (with reverse emitter-base bias)
I_{CM}	Peak collector current or peak pulse current
$I_{C(cont)}$	Continuous collector current
I_D	Drain current
I_{DM}	Pulsed drain current
$I_{D(on)}$	On-state drain current
I_{DSS}	Zero gate voltage drain current
I_E	Emitter current
I_{EBO}	Emitter-base cut-off current
I_F	Forward current
I_{GSS}	Gate-body leakage
I_R	Leakage Current
I_{SD}	Continuous body diode current
I_{USB}	Avalanche current or current under secondary breakdown conditions
I_Z	Zener diode reverse current
N or N_F	Noise figure
N_f	Flicker noise
P_D	Power Dissipation
P_{tot}	Maximum continuous package dissipation
P_{totp}	Practical power dissipation
Q	Figure of Merit (Tuner diodes)
$rbb'C_{bc}$	Collector base time constant
$r_{ce(sat)}$	Collector-emitter saturation resistance
r_d	Series resistance
$R_{DS(on)}$	Drain-source on resistance
R_S	Slope resistance
r_{th} or R_{TH}	Thermal resistance
$R_{TH(trans)}$	Transient thermal resistance
r_Z	Differential resistance
t_d	Delay time
$t_{d(on)}$	Turn-on delay time

Glossary

Symbol	**Parameter**
$t_{d(off)}$	Turn-off delay time
t_f	Fall time
t_{off}	Turn-off time ($t_{off} = t_f + t_{stg}$)
t_{on}	Turn-on time ($t_{on} = t_d + t_r$)
t_r	Rise time
t_{rr}	Reverse recovery time
t_{stg}	Storage time
T_{amb}	Ambient temperature
T_{case}	Case temperature
t_{fr}	Forward recovery time
T_j	Junction temperature
T_{stg}	Storage temperature
V_{BE}	Base-emitter voltage
$V_{BE(on)}$	Base-emitter turn-on voltage
$V_{BE(sat)}$	Base-emitter saturation voltage
$V_{(BR)CBO}$	Collector-base breakdown voltage
$V_{(BR)}$	Breakdown voltage
$V_{(BR)CEO}$ or BV_{CEO}	Collector-emitter breakdown voltage with base open circuit
$V_{(BR)CES}$ or BV_{CES}	Collector-emitter breakdown voltage with base emitter shorted
$V_{(BR)CEV}$ or BV_{CEV}	Collector-emitter breakdown voltage with base emitter reverse biased
$V_{(BR)EBO}$ or BV_{EBO}	Emitter-base breakdown voltage
V_{CB}	Collector-base voltage
$V_{CEO(sus)}$	Collector-emitter sustaining voltage
$V_{CE(sat)}$	Collector-emitter saturation voltage
$V_{GS(th)}$	Gate source threshold voltage
V_R	Reverse Voltage
V_Z	Zener diode breakdown voltage
Z_R	Reverse dynamic impedance
τ	Effective minority carrier lifetime

ZETEX Technical Publications

The following lists the publications available from Zetex. Please contact your regional sales office for complimentary copies. Full addresses and contact numbers appear on the rear cover.

High Performance Surface Mount Devices

This selection guide provides shortform information on Zetex range of high current high efficiency surface mount devices. It includes SuperSOT high performance NPN and PNP bipolar transistors, Darlington transistors, MOSFETs, switching transistors, switching diodes, RF transistors, high voltage transistors, low noise transistors, high performance tuner and Schottky diodes. Packages featured are SOT23, SOT89 and SOT223.

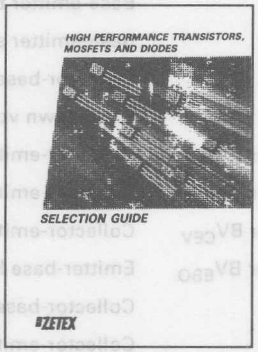
HIGH PERFORMANCE SURFACE MOUNT DEVICES

SELECTION GUIDE

ZETEX

High Performance Transistors, MOSFETs and Diodes

This selection guide provides shortform information on Zetex through hole products. It includes high performance NPN and PNP bipolar transistors, Darlington transistors, MOSFETs, switching transistors, IGBTs, switching diodes, RF transistors, low noise transistors and tuner diodes. These products feature the Zetex E-Line (TO-92 style) package.

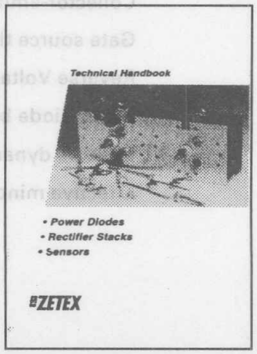

HIGH PERFORMANCE TRANSISTORS, MOSFETS AND DIODES

SELECTION GUIDE

ZETEX

Technical Handbook; Power Diodes, Rectifier Stacks and Sensors.

This databook provides full technical information on Rectifier diodes, Rectifier Stacks, Power Zener/Suppressor diodes, Magnetic Field Sensors and Current Sensors.


Technical Handbook

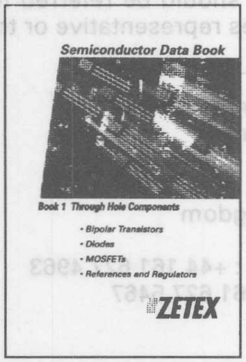
- Power Diodes
- Rectifier Stacks
- Sensors

ZETEX

ZETEX Technical Publications

Discrete Through Hole Components Data Book

This data book provides full information on Zetex through hole products. It includes high performance NPN and PNP bipolar transistors, Darlington transistors, MOSFETs, switching transistors, IGBTs, switching diodes, RF transistors, low noise transistors and tuner diodes. These products feature the Zetex E-Line package as well as the industry standard TO92.



Semiconductor Data Book

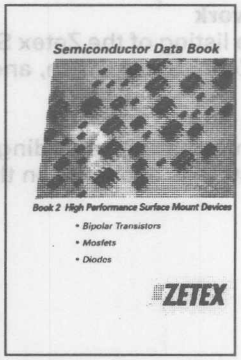
Book 1 Through Hole Components

- *Bipolar Transistors*
- *Diodes*
- *MOSFETs*
- *References and Regulators*

ZETEX

Discrete Surface Mount Components Data Book

This data book provides full information on the Zetex range of surface mount products. It includes high performance bipolar products, Darlingtons, MOSFETs, IGBTs, switching and RF transistors, variable capacitance, switching and zener diodes. Package options include SOT23, SOT89, SOT223 and SM8 (dual transistor SOT223).



Semiconductor Data Book

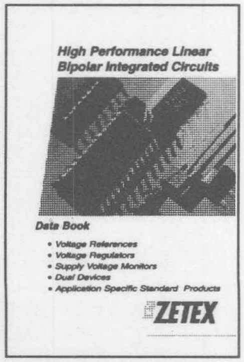
Book 2 High Performance Surface Mount Devices

- *Bipolar Transistors*
- *Mosfets*
- *Diodes*

ZETEX

High Performance Linear Bipolar Integrated Circuits

This data book provides full information on the Zetex range of integrated circuits. It includes voltage references, regulators, supply monitors, dual devices and application specific standard products. Both through hole and surface mount packages are featured.



High Performance Linear Bipolar Integrated Circuits

Data Book

- *Voltage References*
- *Voltage Regulators*
- *Supply Voltage Monitors*
- *Dual Devices*
- *Application Specific Standard Products*

ZETEX

ZETEX Technical Services

Applications/Technical Assistance :

Requests for additional technical information, sample availability or applications assistance should be referred to one of the following; your regional Zetex Sales Office, sales representative or the Applications Support Group at :-

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Fields New Road
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Oldham
OL9 8NP
United Kingdom

Telephone : +44 161 627 4963
Fax : +44 161 627 5467

Sales Network
A complete listing of the Zetex Sales Locations, Distributors and Agents is available from any Zetex Sales Office, and on the Zetex Web Site.

Zetex Web Site
Updated information regarding Zetex, the product range, latest news and sales network listing is available on the Zetex Web Site at http://www.zetex.com

Section 1

Application Notes

ZETEX

Application Note 1
Issue 2 October 1995

8 Watt Fluorescent Lamp Inverter

Compact DC to AC Inversion for Efficient Lighting Systems

David Bradbury

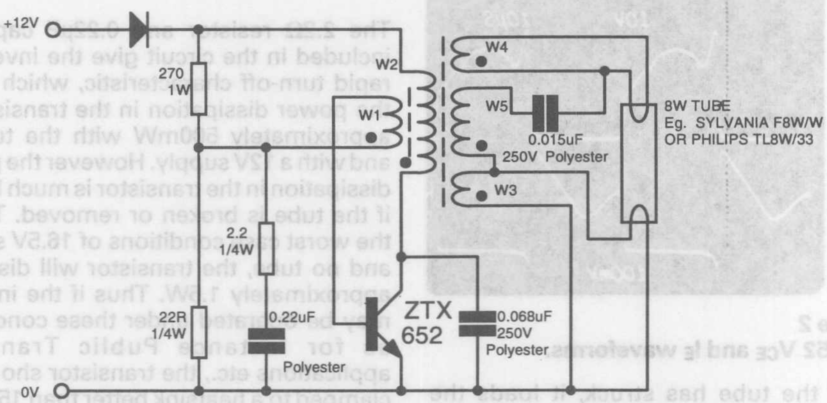


Figure 1
8W Invertor.

The circuit shown in Figure 1 has been designed to drive an 8W fluorescent lamp from a 12V source, using an inexpensive inverter based on the ZTX652 transistor. The inverter will operate from supplies in the range of 10V to 16.5V, attaining efficiencies up to 78% thus making it suitable for use in on-charge systems such as caravans / mobile homes/ RVs as well as periodically charged systems such as roadside lamps, camping lights or outhouse lights etc. Other features of the inverter are that it oscillates at an inaudible 20kHz and that it includes reverse polarity protection.

Circuit Operation

The 270Ω and 22Ω resistors bias a ZTX652 transistor into conduction, where the positive feedback given to the transistor by W1 drives it into saturation, thus applying the supply voltage across W2. This causes a magnetising current to build up in W2 until the transformer's ferrite core saturates. When this happens, the base drive given to the transistor by W1 decays, causing the transistor to rapidly turn off.

Until the fluorescent tube strikes, the transformer is only loaded by the tube heater filaments which present only a minimal load. Thus when the transistor turns off the transformer 'rings' for half a cycle at a frequency governed by the windings' inductance and the 0.068µF capacitor, reversing the magnetising current and turning the transistor on again. This 'ring' induces a high voltage pulse across the fluorescent tube which will cause it to strike once the heaters have warmed up.

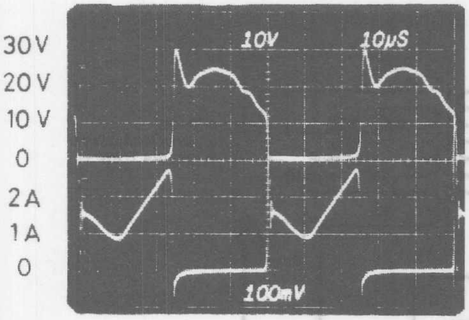


Figure 2
ZTX652 V_{CE} and I_E waveforms.

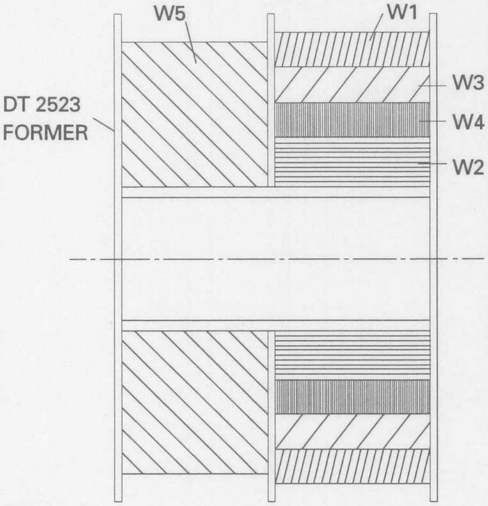
Once the tube has struck, it loads the transformer heavily, swamping this ringing action and so greatly reducing the peak voltage induced across W2 and the transistor. This extends the non-conducting period of the inverter cycle and during this period, energy stored in the transformer in the form of magnetising current is dumped into the fluorescent tube. When all of this energy has been dumped, the voltage on the transistor collector falls and it switches on once more.

The voltage required to pass current through the tube has now fallen enough for it to conduct during both half cycles of inverter oscillation. Thus when the transistor now turns on, it both drives the fluorescent tube directly and also stores energy in the transformer which drives the tube during the transistor's non-conducting period. The current passed through the tube is controlled by the transformer's leakage inductance and also a series connected 0.015µF capacitor. Waveforms of the transistor's collector voltage and emitter current under normal operating conditions are given in Figure 2.

The 2.2Ω resistor and 0.22µF capacitor included in the circuit give the inverter a rapid turn-off characteristic, which limits the power dissipation in the transistor to approximately 500mW with the tube lit and with a 12V supply. However the power dissipation in the transistor is much higher if the tube is broken or removed. Taking the worst case conditions of 16.5V supply and no tube, the transistor will dissipate approximately 1.5W. Thus if the inverter may be operated under these conditions as for instance Public Transport applications etc., the transistor should be clamped to a heatsink better than 15 °C/W.

Where the inverter will not remain energised if the tube does not strike, no heatsink is necessary.

Similar higher power designs are possible using the ZTX853 4A DC rated part, which is available in the E-Line package, (enhanced TO92) and also in the surface mount SOT223 package as the FZT853.



Transformer details. Core type FX3439 with 0.005 in (0.125 mm) spacer		
Former type DT2523		
Winding	Turns	Material
W1	4	34swg. Enamelled Copper Wire
W2	17	26swg. Enamelled Copper Wire
W3	7	28swg. Enamelled Copper Wire
W4	7	28swg. Enamelled Copper Wire
W5	130	36swg. Enamelled Copper Wire

Figure 3
Transformer winding detail.

Construction

Apart from keeping component lead lengths short, the layout of the circuit is not critical. Care should be taken however in winding the transformer (see figure 3). It is advisable to mount the unit in a metal case as this will provide RF screening of the inverter and also provides a ground plane for the fluorescent tube which significantly reduces its striking voltage. The case could also be used as heatsinking for the ZTX652 transistor when required.

Notes

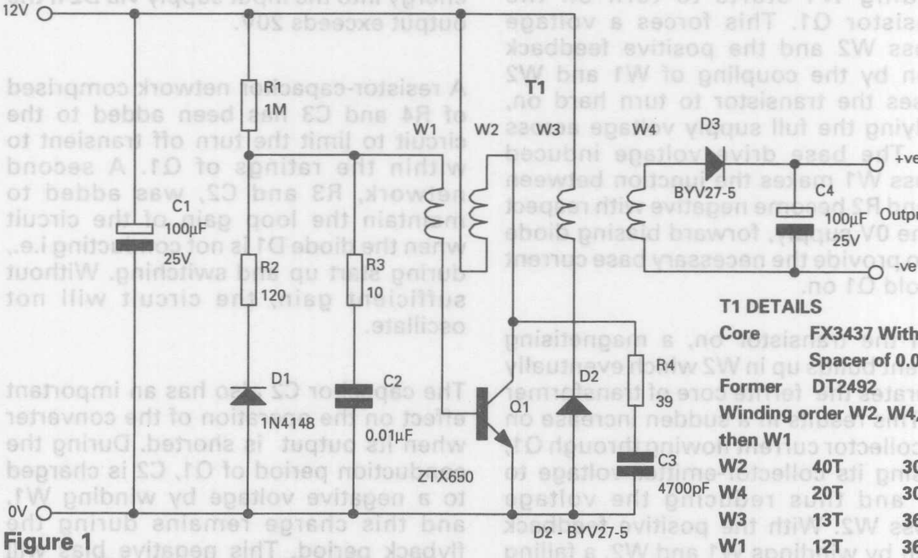
Portable Ni-Cd Battery Charger

Inexpensive Load Tolerant Step-Up Energy Source

David Bradbury

Many portable systems such as radio/tape players and video recording equipment are commonly powered by a 12V rechargeable Nickel-Cadmium (Ni-Cd)battery pack, which rarely has the capacity to power the system for more than a few hours. It would be convenient if these batteries could be recharged from a readily available source such as an automotive battery.

The circuit shown in Figure 1 was designed to charge 12V, 2 ampere-hour Ni-Cd battery packs from a 12V source, but was made versatile enough to charge packs in the range of 4.8 to 15.6V to increase possible applications. To charge batteries that may reach a higher end voltage than the 12V supply input (even 12V Ni-Cd packs reach an end voltage of about 14.5V), a voltage



T1 DETAILS

Core	**FX3437 With Gap/ Spacer of 0.08mm**	
Former	**DT2492**	
Winding order W2, W4, W3 then W1		
W2	**40T**	**30swg**
W4	**20T**	**30swg**
W3	**13T**	**36swg**
W1	**12T**	**36swg**

Figure 1
Flyback Topology based Ni-Cd Charger.

converter of some kind is required. This converter must be short and open circuit protected, be able to operate efficiently over a wide output voltage range and must be insensitive to input voltage variations.

A self oscillating flyback converter will meet most of these requirements, being particularly suited to operating over a wide output voltage range. However, the standard circuit needs modifications to cope satisfactorily with short and open circuit loads. These changes include a special biasing circuit and an energy recovery winding on the converter transformer.

Circuit Operation

When power is first applied to the circuit, a small bias current supplied by R1 via winding W1 starts to turn on the transistor Q1. This forces a voltage across W2 and the positive feedback given by the coupling of W1 and W2 causes the transistor to turn hard on, applying the full supply voltage across W2. The base drive voltage induced across W1 makes the junction between R1 and R2 become negative with respect to the 0V supply, forward biasing diode D1 to provide the necessary base current to hold Q1 on.

With the transistor on, a magnetising current builds up in W2 which eventually saturates the ferrite core of transformer T1. This results in a sudden increase on the collector current flowing through Q1, causing its collector-emitter voltage to rise and thus reducing the voltage across W2. With the positive feedback given by windings W1 and W2, a falling voltage across W2 causes Q1 to turn off rapidly.

The current flowing in W2 now forces the collector voltage of Q1 to swing positive until restricted by transformer output loading. During this “flyback” period, the voltage induced across W4 forward biases the diode D3 to charge the output capacitor C4. Energy stored in the core of the transformer while the transistor was on is dumped into this capacitor which feeds the load. The collector voltage of Q1 remains high until the current flowing in W4 falls to zero.The voltage then falls until the positive feedback given by W1 causes Q1 to turn on hard again to start the next cycle of oscillation. If a load (Ni-Cd battery) is not connected across C4, the energy dumped into this capacitor will charge it to an ever increasing voltage. To restrict the maximum output voltage, an extra transformer winding, W3, has been added which will return stored energy into the input supply via D2 if the output exceeds 20V.

A resistor-capacitor network comprised of R4 and C3 has been added to the circuit to limit the turn off transient to within the ratings of Q1. A second network, R3 and C2, was added to maintain the loop gain of the circuit when the diode D1 is not conducting i.e., during start up and switching. Without sufficient gain, the circuit will not oscillate.

The capacitor C2 also has an important effect on the operation of the converter when its output is shorted. During the conduction period of Q1, C2 is charged to a negative voltage by winding W1, and this charge remains during the flyback period. This negative bias will inhibit continuous oscillation unless the transformer “rings” sufficiently at the end of the flyback period to produce a transient base drive voltage large enough to overcome the bias. Since an output voltage of at least 1.5V is required to produce sufficient transformer ringing, a short circuit load causes the converter to run in an intermittent mode, consuming very little power.

Converter Design

The converter is to charge a 12V, 2Ah battery pack, which has a recommended charge current rating of 220mA. The typical power source is a 12V car battery.

Firstly, a transformer core must be chosen that will give the necessary throughput without the need to operate at an excessive frequency. The choice will be controlled by the peak current passed through winding W2 and the inductance required. The efficiency of the converter can be expected to be around 75%, so with a 12V supply and 12V output load, the average supply current will be:

$$I_s = \frac{I_o \times V_o}{V_s \times E_{ff}} = \frac{0.22}{0.75} = 0.29\text{A}$$

The actual supply current taken by the converter will be a linear ramp from zero to I_{peak} followed by a period of no current flow. The ratio of the ramp period to the whole cycle period is the duty cycle. Because of the simple current waveform, once the duty cycle is known, the peak current in W2 can be calculated from the average supply current. The duty cycle is dependent on the input to output turns ratio of the transformer. The smaller the number of turns on the output winding, the higher the flyback voltage across the switching transistor for a given output voltage, resulting in a shorter flyback period. Reducing the flyback period allows a given output power to be achieved with a small peak current in the switching transistor, helping to minimise losses.

However this is at the expense of requiring a higher voltage transistor. A compromise duty cycle of 70% was chosen for this design. This gives I_{peak} as:

$$I_{peak} = \frac{I_s \times 2}{\text{Duty } C_{y.}} = \frac{0.29 \times 2}{0.75} = 0.83\text{A}$$

The E-Line ZTX650 transistor will yield a high gain at this current and so was chosen as the switching transistor. To keep the converter inaudible yet minimize switching losses, an oscillation frequency in the range of 20 to 50kHz was chosen. This gives a transistor on time (current build up time) of 35μs to 14μs respectively.The inductance of transformer winding W2 can now be calculated using:

$$L_{max} = \frac{V_s \times T_{on}}{I_{peak}} = \frac{12 \times 35E{-}6}{0.83} = 0.5\text{ mH}$$

Similarly, L_{min} was calculated to be approximately 0.2mH.

The energy storage capability of the suitable RM range of transformer cores are described in the form of Hanna curves. These curves relate $I^2 \times L$, $I \times N$ and core spacer. The $I^2 \times L$ value that is required for this transformer is 0.33E-3 to 0.17E-3. The smallest core in the RM range will meet this specification. An RM type FX3437 pair of cores with a 0.08mm spacer will give an $I^2 \times L$ factor of 0.25E-3. This factor is in the required range and also corresponds to a pre-gapped RM core type LA14376 which thus can be used as a convenient substitute. The inductance of W2 will be $(0.25E\text{-}3/(0.83^2)) = 0.36mH$, requiring 38.5 turns according to the Hanna curves for this core and gap (rounded to 40 turns).

The output winding W4 is determined by:

$$W_4 = \frac{V_o \times T_{off} \times W_2}{V_S \times T_{on}} = \frac{12.7 \times 30\% \times 40}{12 \times 70\%}$$

= 18.4, rounded to 20

Note the ratio of T_{on} to T_{off} was used above.

The output voltage must be limited to 20V by winding W3 and so this gives:

$$W_4 = \frac{(V_s + V_{REC}) \times W_4}{(V_{max} + V_{rec})} = \frac{12.7 \times 20}{20.7}$$

= 12.2, rounded to 13

The base winding W1, is a compromise between providing sufficient base current for a low gain transistor operating with minimum supplies, and avoiding losses caused by overdriving the transistor under normal circumstances. The transistor is required to pass 0.83A peak and a minimum gain device at low temperature will need approximately 15mA to achieve this. A base drive voltage of at least 1.4V is needed to pass any current at all through the biasing circuit adopted and a voltage of twice this is desirable if the base current is to be insensitive to supply voltage variations. A base winding of 12 turns will give a drive of 3.3V with a minimum supply of 10.5V. A base resistor of 120Ω gives the required base current from this drive.

Finally, the starting base bias resistor value must be calculated. This resistor must cause the circuit to have sufficient gain to oscillate yet not cause excessive power dissipation if oscillation does not occur due to incorrect winding phasing or some other fault. To oscillate, the loop gain of the circuit must be greater than one. The feedback gain is 12/40 or about 0.3, so the transistor must give a voltage gain of at least 1/0.3 or 3.3. The transistor voltage gain is mainly dependent on the ratio of collector and emitter loads. The small signal collector loading is the result of the tuned circuits made by the transformer windings and associated capacitors, turning out to be of the order of 2k. The emitter loading is the intrinsic emitter resistance, given by $r_e = 25E\text{-}3/I_e$. For a voltage gain of 3.3, r_e must be less than 2k/3.3 or 600. The minimum emitter current during start up must be more than 25E-3/600 or 42μA. The h_{FE} of the ZTX650 transistor is not specified at such low currents but it is not expected to be less than 30, so this would set the minimum base current at 1.5μA. However, because of this uncertainty, the bias was raised to 10μA to ensure reliable starting. This value will only cause a worst case power dissipation of about 70mW in the transistor if the circuit fails to oscillate under fault conditions.

Performance

Over the intended operational range, the circuit was found to give an efficiency exceeding 70%, providing a useful output from a supply as low as 9V. Full input and output characteristics of the converter are given in Figures 2 and 3. Figure 2 shows the output current given by the circuit for various load voltages. Note the output current at 12V is very close to the design aims. This diagram also shows the efficiency of the converter when operating into these loads. Figure 3 shows how the output current given into a 12V load varies with input supply voltage.

The time required to fully charge the load batteries will depend on their voltage and ampere-hour capacity. The converter was designed to charge the 2Ah power pack in about 14 hours. At this charge rate, these vented battery packs will safely stand continuous over charging. If the converter is used to charge a different battery pack, Figure 2 should be used to find the output current of the circuit which can then be used to calculate the charge time necessary. In cases where the charge rate is greater than C/10 for vented cells or C/50 for button cells, it is recommended that a timer be included in the circuit to ensure that accidental overcharging does not occur.

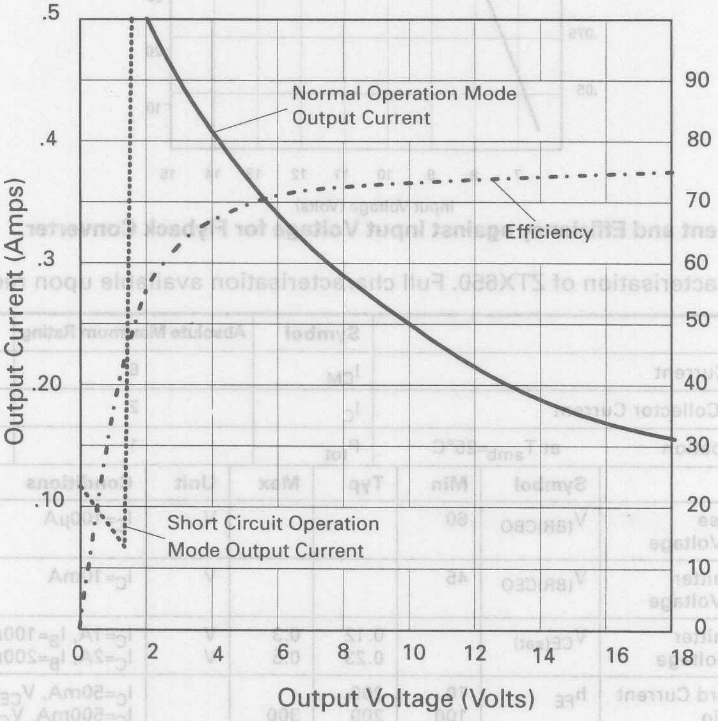


Figure 2
Output Current and Efficiency against Output Voltage for Flyback Converter.

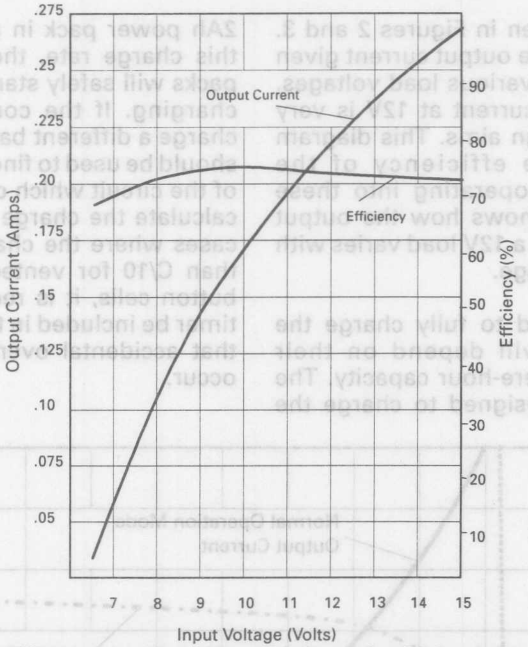


Figure 3
Output Current and Efficiency against Input Voltage for Flyback Converter.

Partial Characterisation of ZTX650. Full characterisation available upon request.

Parameter		Symbol	Absolute Maximum Rating	Unit
Peak Pulse Current		I_{CM}	6	A
Continuous Collector Current		I_C	2	A
Power Dissipation	at T_{amb}=25°C	P_{tot}	1	W

Parameter	Symbol	Min	Typ	Max	Unit	Conditions
Collector-Base Breakdown Voltage	$V_{(BR)CBO}$	60			V	I_C=100μA
Collector-Emitter Breakdown Voltage	$V_{(BR)CEO}$	45			V	I_C=10mA
Collector-Emitter Saturation Voltage	$V_{CE(sat)}$		0.12	0.3	V	I_C=1A, I_B=100mA
			0.23	0.5	V	I_C=2A, I_B=200mA
Static Forward Current Transfer Ratio	h_{FE}	70	200			I_C=50mA, V_{CE}=2V
		100	200	300		I_C=500mA, V_{CE}=2V
		80	170			I_C=1A, V_{CE}=2V
		40	80			I_C=2A, V_{CE}=2V

ZETEX

Application Note 3
Issue 2 November 1995

Infra-Red Remote Control and Data Transmission

An Introduction to Photodiodes - Load Circuits and Applications

David Bradbury

Introduction

The use of short range remote control and data transmission systems in both consumer and industrial products is growing rapidly. Already common-place in items such as Televisions, Teletext and Viewdata controllers, toy model control etc, these systems are now being used for the remote control of High Fidelity Units, Garage Doors, Light Dimmers, Slide Projectors, and in Teaching Aids, Data links, Burglar alarms etc.

In the past, short range remote control systems have mainly relied on Radio or Ultrasonic links to effect control. However, these methods have serious disadvantages. Radio links often require licensing, they can give excessive range and are prone to electrical interference. Ultrasonic links suffer from multipath interference which limits their useable data range, and interference from numerous everyday objects that produce sound at ultrasonic frequencies, such as keys, coins, bells, electrical apparatus etc. Infra-Red (IR) light links can be used for short range remote control and their freedom from many of these problems, in part, explains their rapidly growing popularity.

Generally IR links consist of a modulation source driving a light emitting diode that radiates at a wavelength of 850 to 970nm. This light is detected by a photodiode, and the resulting signal is amplified and decoded to recover the transmitted information. Since IR light is used, licensing is not required; the radiation is easily confined to a single room since walls and doors block these wavelengths, and electrical interference is easily rejected. Also, multipath interference does not significantly degrade the signal, and there are few domestic light sources emitting IR that flicker at a frequency high enough to corrupt a modulated signal. (Compact Fluorescent or "energy saving" lamps are a possible problem, though manufacturers do try to avoid common IR data transmission frequencies).

The main limitations of an optical system result from the low power output available from an IR light emitting diode (LED), combined with noise generated in the photodiode by current flowing in the device due to ambient lighting and leakage. These factors control the operational range of a system and the ambient light levels it can tolerate.

The transmitted power can effectively be increased by using lenses to concentrate the IR LED's light output, but this narrows the transmission beam width making alignment too critical for some applications. The receiver signal to noise ratio can be optimised by using a very low leakage, high sensitivity photodiode in conjunction with an optical filter, which only passes wavelengths emitted by IR light emitting diodes.

The Zetex BPW41D photodiode has been developed specifically for use in IR links. It is a high speed, low capacitance device enclosed in a package which acts as a highly selective IR pass filter.

Infra-Red Photodiodes

The Zetex BPW41D photodiode, is a low leakage silicon p.i.n. (p-intrinsic-n) device of planar construction and possesses an active area of 7.5 mm^2. The device has a silicon nitride layer over the chip which acts both as a passivating and an efficient anti-reflection coating. The plastic housing of the BPW41D contains a dye which transmits well in the near infra-red part of the spectrum (800 nm to 1100 nm) but which strongly absorbs visible light (400 nm to 700 nm). The insensitivity of the device to visible light and it's response to infra-red radiation are aided by the silicon spectral response, which is low in the blue-green regions and high in the infra-red.

The BPW41 die is also available in a clear encapsulated package as the BPW41C. A smaller die geometry version of the device, the ZPD200, is ideal for short range and low cost IR systems.

Planar construction is used to keep the reverse leakage current low which is important in small signal applications. The p.i.n. structure gives two main advantages. Firstly, capacitance per unit area is lower than that of a conventional p.n. diode, leading to a higher speed of response. Secondly, the minority carrier lifetime is, in general, higher than in heavily doped silicon giving somewhat greater IR sensitivity. (Please refer to Appendix for selected BPW41D data).

Photodiode Load Circuits

[Although originally produced with respect to the BPW41, the following comments and circuits are equally applicable to other photodiodes such as the ZPD200, with minimal or no change to component values].

When used in typical remote control links, the BPW41D can provide useable signal to noise ratios with signal levels that can only produce photocurrents of the order of 10nA. However, to detect such levels, the load used for the photodiode must be carefully chosen to suit the particular operating conditions.

Simple Resistive Loads

The simplest load that can be used is a resistor as shown in Figure 1. In this circuit the photodiode is connected reverse biased across the power supply with a resistor R in series. Any signal current produced by the photodiode will develop a voltage across R which can then be amplified to give the required output.

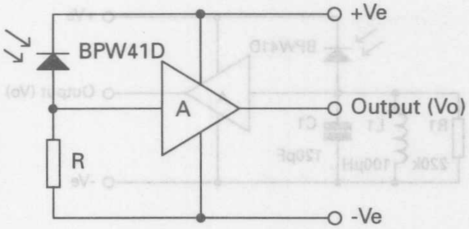


Figure 1
Simple resistive load photodiode detection circuit.

There are various constraints in choosing the value of R. The lower R is made, the smaller will be the signal voltage appearing across it. For a given output voltage this means the gain of amplifier A will have to be higher, making it's noise contribution more significant, and interference rejection and stability harder to achieve. As a result, if R is made too small the operating range of the IR link is reduced.

With a high input impedance amplifier the bandwidth of the circuit will be controlled by the capacitance of the photodiode and the value of R. As R is increased the cut-off frequency falls. Since most IR links use short duration light pulses for control, the receiver circuit must have sufficient bandwidth to detect these pulses without significant attenuation. This gives one limitation on the maximum value of R.

A second and possibly more stringent limitation to the maximum value of the load resistor, is caused by photocurrent in the BPW41D due to ambient lighting. Although the BPW41D includes a selective optical filter to reject light below 700nm, many light sources give significant proportion of their energy output above these wavelengths. The effect of these sources on the photodiode is to cause a steady d.c. current to flow through the device, developing a voltage across R.

This reduces the reverse bias applied to the photodiode, increasing it's capacitance and so reducing the bandwidth. Also, if R is much too large, the voltage developed by ambient light photocurrent can exceed the supply voltage, with the photodiode becoming forward biased. In this condition any signal current detected will be dissipated in the forward biased shunt resistance of the photodiode, and so the signal output level will be severely attenuated.

As a result of these problems the simple resistor load is mainly used in applications where ambient light levels are low. For instance, it is sometimes used in the remote control of television receivers where it's obvious simplicity keeps cost down, and it's inability of operating efficiently in high light levels is unimportant as the television is unlikely to be operated under such conditions. The values of R typically used in remote control systems range from 100kΩ to 300kΩ giving circuit cut-off frequencies of 30kHz to 100kHz and an ability to handle ambient light levels that cause 30µA to 100µA of photocurrent. This corresponds to a reasonably well lit room using tungsten filament lamps.

The simple resistor load can be modified to handle higher light levels without a large sensitivity loss with the circuit shown in Figure 2.

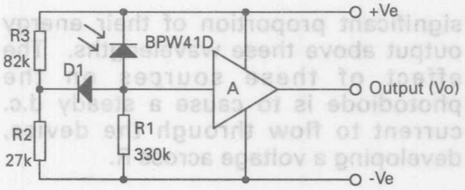


Figure 2
Modified resistive load photodiode circuit to allow high ambient light levels.

At low ambient light levels this circuit will operate with the 300kΩ resistor as the photodiode load, giving a similar sensitivity to the simple load. At high light levels the voltage across the 300kΩ resistor exceeds the voltage set up by the potential divider R2, R3. This causes diode D1 to conduct so that the effective load of the photodiode is the potential divider circuit which has a resistance of approximately 20kΩ. This gives a x15 loss in sensitivity, but the circuit will continue to work in ambient light levels approaching direct sunlight. If the photodiode is not prevented from becoming forward biased by this, or alternative circuits, the loss in sensitivity at high levels will be in the order of ×1000.

Inductive Loads

The optimum load for a photocell in a remote control circuit has a high impedance at signal frequencies and a low impedance at any other frequencies including d.c. These load characteristics can be achieved with an inductive load.

The inductive load shown in Figure 3 can provide a signal frequency impedance of 100kΩ whilst still giving a very low d.c. current resistance path for photocurrent caused by ambient lighting. As a result, it will operate well over a wide range of ambient light levels. The inductor is normally tuned to match the input frequency, and damped with a resistor to suppress ringing.

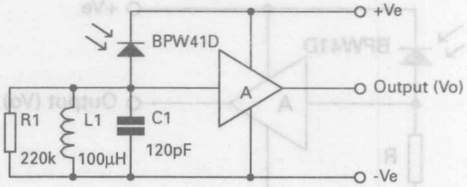


Figure 3
Inductive load photodiode detection circuit.

The output of the circuit when pulsed with IR light is shown in Figure 4. It consists of a damped sine wave that takes several cycles to decay. Unfortunately this ringing reduces the maximum data rate the circuit will support, since delays must be included in the receiver decoder to avoid detecting multiple pulses as the sine

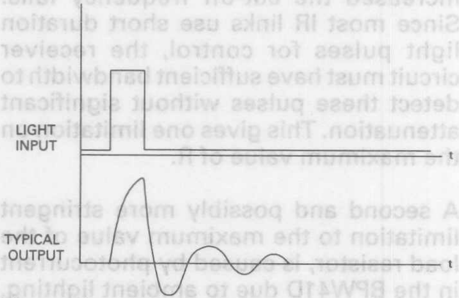


Figure 4
Output response of inductive load photodiode detection circuit.

wave decays. Some ringing is unavoidable due to the capacitance of the photodiode and other stray effects acting with the inductive load, so that the best solution is to tune the load to match the signal, and dampen with a resistor that does not significantly reduce the impedance of the load.

The circuit does have other disadvantages which limit it's usefulness. It is difficult to wind the inductor without it's natural self capacitance bringing the self resonant frequency below the operational frequency required. To keep it's mechanical size small, it is necessary to use a ferrite core to reduce the number of turns in the inductor. However d.c. current flowing in the inductor due to ambient light photocurrent can saturate the ferrite core if the light level is high and the core volume small, thereby reducing the inductance of the coil and lowering the impedance of the load at signal frequencies. This problem adds further restrictions on the design of the inductor.

The inductive load circuit is used in the remote control of television receivers as an alternative to the simple resistive load. It can give a sensitivity similar to that of the resistive load, has better high ambient light level tolerance, but a lower data transmission rate.

Active Loads

The optimum load characteristic of high impedance at high frequencies, yet low impedance at low frequencies and d.c., can be achieved with an active load. Three configurations of the same basic active load circuit, but with differing output polarities and output impedances are given in Figures 5, 6 and 7.

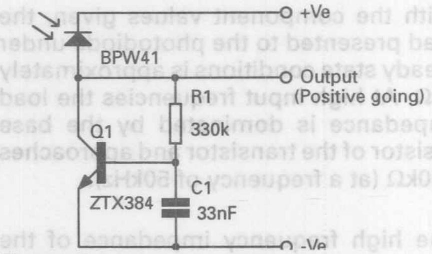


Figure 5
Active load photodiode circuit -1.

Consider the circuit shown in Figure 5. In this circuit, current passed by the BPW41D flows through the 330kΩ resistor to the base of ZTX384 transistor. This causes the transistor to conduct, shunting the photodiode current directly to the negative supply and so reducing the base drive to the transistor. An equilibrium point is reached in which the transistor holds it's collector voltage at approximately 0.8V by acting as a constant current generator that exactly matches current fed to it by the photodiode. This equilibrium is maintained for d.c. or slowly varying photocurrents, so giving the photodiode a very low impedance load under these conditions.

The current dump transistor has a capacitor connected to it's base, which restricts the speed at which the load circuit can respond to a sudden change in photocurrent. As a result, the current matching equilibrium of the load circuit is not maintained for rapidly changing photocurrents. This makes the high frequency impedance of the load very much higher than it's low frequency impedance.

With the component values given, the load presented to the photodiode under steady state conditions is approximately 1kΩ. At high input frequencies the load impedance is dominated by the base resistor of the transistor and approaches 250kΩ (at a frequency of 50kHz).

The high frequency impedance of the load drops slightly under high ambient light levels, but the main disadvantages of the circuit shown in Figure 5 are it's noise contribution and susceptibility to interference at high light levels. Both of these problems stem from the extremely high amplification given to any signal generated in the base-emitter circuit of the load transistor under high ambient lighting conditions.

The voltage gain given to a signal generated in the base emitter circuit is approximately:

Gain $\approx R_c/r_e$,

where R_c is the collector load impedance i.e. 250kΩ, and r_e is the intrinsic emitter resistance. $r_e = 25\ \Omega/I_e$ where I_e is the transistor emitter current (in mA).

i.e. Gain $\approx R_c \times I_e/25$ (I_e in mA)

If ambient lighting causes a photocurrent of 5μA to flow through the load (Eg.current through the photodiode in a dimly lit room), then:

Gain $\approx 250 \times 10^3 \times 5 \times 10^{-3}/25 = 50$ for 5μA photocurrent.

However if the ambient light level approaches direct sunlight:

Gain $\approx 250 \times 10^3 \times 1/25 = 10{,}000$ for 1mA photocurrent.

This very high amplification given to noise generated in the load transistor degrades the performance of this load circuit when operated at high ambient light levels.

The gain at high photocurrents can be significantly reduced by including a resistor in the emitter of the current dump transistor as shown in Figure 6. The performance at low light levels is unaffected, but the voltage gain given to noise signals in the base emitter circuit of the load is much lower when ambient light levels are high.

$$\text{Gain} \approx R_c/(r_e + R_e)$$
$$= 250 \times 10^3/(25+1.5 \times 10^3)$$

≈ 160 for 1mA photocurrent.

The added resistor restricts the maximum gain given by the load transistor, so limiting the noise and interference introduced by the circuit.

The emitter resistor also increases the impedance of the load at low frequencies, and so it's value is a compromise between minimum noise contribution and minimum d.c. voltage drop at high photocurrents. The value used in Figure 6 ensures that the noise generated by the load is less than that of the photodiode, yet the circuit can still operate in direct sunlight.

To take advantage of the high impedance of these load circuits, the amplifier connected across the load must have a high input impedance.

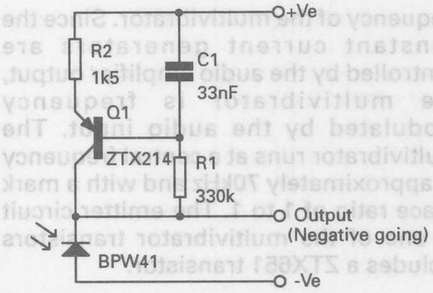


Figure 6
Active load photodiode circuit -2.

If a low impedance amplifier is to be used, an emitter follower buffer can match the photocell load to the amplifier as in Figure 7. The emitter follower is d.c. coupled to the load to eliminate the coupling and biasing components that otherwise would have been required. It has an output impedance of approximately 2kΩ at small signal levels.

All three loads may be modified to give signal outputs of opposite polarity simply by exchanging ZTX214C transistors for ZTX384C and vice versa, and reversing the power supply connections and those of the BPW41D.

Remote Control using the BPW41D and similar Photodiodes

It is a vital requirement that remote control circuits cannot be triggered by spurious signals. To achieve this immunity without sacrificing range of operation, coded transmissions are normally used so that only received signals of a pre-determined sequence are allowed to alter control outputs. Effective coding and decoding requires elaborate circuitry if constructed using standard logic devices. However, numerous dedicated integrated circuits have been developed for this task. These systems make use of separate encoder and decoder ICs. The encoder performs a scanning function on a matrix configured keypad, and on detecting a key press, generates a corresponding coded string. The string is repeated until the key is released.

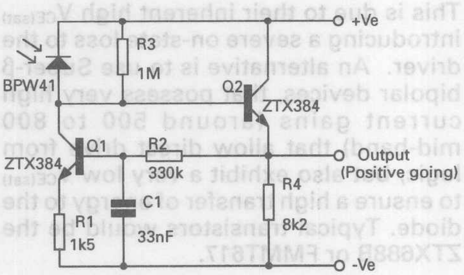


Figure 7
Active load photodiode circuit - 3.

The output of the IC typically drives a bipolar transistor that in turn drives either a single, or an array of IR light emitting diodes. To maximise range these diodes are pulsed at 1 to 2A, but as the duty cycle is so low the average current drain on the battery is low. It is important to pay attention to PCB layout to maintain a low resistance and low inductance path for the diode/driver/supply. To allow higher battery capacity/volume, and hence longer lifetime, means that these systems are more efficient with lower voltage battery packs, typically two cells (3V). In these cases, and to allow the IR LED to be pulsed effectively from logic IC outputs, the use of Darlington transistors is excluded.

This is due to their inherent high $V_{CE(sat)}$ introducing a severe on-state loss to the driver. An alternative is to use Super-β bipolar devices, that possess very high current gains (around 500 to 800 mid-band) that allow direct drive from logic, but also exhibit a very low $V_{CE(sat)}$ to ensure a high transfer of energy to the diode. Typical transistors would be the ZTX688B or FMMT617.

Monophonic Audio Link (M.A.L.)

An infra-red link capable of transferring audio signals has been constructed using the BPW41D photodiode as the receiving element. The link was designed to allow the use of monophonic headphones without the inconvenience of trailing leads. It uses a frequency modulated system to ensure that volume levels are independent of range. Depending on the ambient light levels and reflectivity of room walls, the link gives satisfactory performance up to a range of 8m. It's frequency response is from 50 Hz to 8kHz.

M.A.L. Transmitter

(Refer to Figure 8.) The audio signal to be transmitted is fed through a gain control R1 to an audio amplifier which gives the circuit an input sensitivity of 100mV RMS. As part of a noise reduction system, the audio amplifier boosts signals at high frequencies by applying 55µs pre-emphasis.

The output of the audio amplifier is capacitively coupled to two constant current generators, these being used to charge the timing capacitors of a multivibrator. Consequently, the generators control the oscillation frequency of the multivibrator. Since the constant current generators are controlled by the audio amplifier output, the multivibrator is frequency modulated by the audio input. The multivibrator runs at a central frequency of approximately 70kHz and with a mark space ratio of 1 to 1. The emitter circuit of one of the multivibrator transistors includes a ZTX651 transistor.

The ZTX651 transistor is used as a buffer to provide the high current drive necessary for the infra-red light emitting diodes, (IR LEDs). Six emitters wired in two groups of three provide the modulated light output, their peak drive current being limited to 300mA.

It is possible for simple multivibrators to attain a latch-up state in which both transistors conduct continuously in saturation, giving a loop gain too low for oscillation. In this circuit the supply for the constant current generators which bias the multivibrator transistors is derived from the collector loads of the multivibrator. As a result, if latch-up should occur, the base drive for the multivibrator will decay, bringing the transistors out of saturation, so raising the loop gain and enabling normal oscillation to occur.

The supply for the current generators is regulated by a zener diode to minimise the frequency variations resulting from power supply voltage fluctuations. The regulating zener also has the effect of restricting the multivibrator collector voltage swing to approximately 6V. This prevents the charge on the timing capacitors reaching V_{BE} breakdown levels and subsequent damage of the

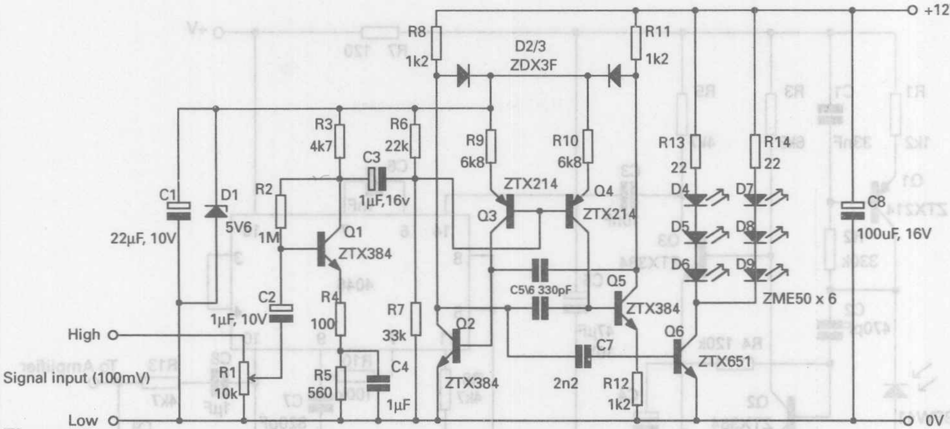


Figure 8
M.A.L. Transmitter unit.

emitter-base junctions of the multivibrator transistors. The supply for the audio amplifier is also taken from this regulator to restrict it's output swing to that required by the current generators. Construction of the transmitter is not critical.

M.A.L. Receiver

(Refer to Figure 9.) The receiver uses a BPW41D photodiode to detect the infra-red signal emitted by the transmitter. The photodiode has an active load which reduces the effects of ambient lighting. The signal voltage appearing across the load is coupled into a very high gain amplifier which boosts it to a suitable level for F.M. demodulation.

Frequency demodulation is achieved using a standard CMOS phase locked loop (pll) logic device. The output of the amplifier is a.c. coupled into a self biasing Schmitt Trigger input in the phase locked loop. The signal is compared with an internal voltage controlled oscillator by a phase comparator, which produces an error voltage that is dependent on the phase difference. This error signal is used to adjust the frequency of the voltage controlled oscillator so that it is locked, or made identical to the incoming signal frequency. When the signal is locked, the voltage output of the phase comparator is proportional to the incoming signal frequency, and so reproduces the audio signal fed into the transmitter.

The output of the phase detector is fed through a de-emphasis network which has a time constant of 55µs. This has the effect of reducing the noise and signal amplitudes at high frequencies, since the signal at the transmitter was boosted by an identical network. The pre-emphasis and de-emphasis method gives a flat link frequency response whilst significantly reducing receiver noise.

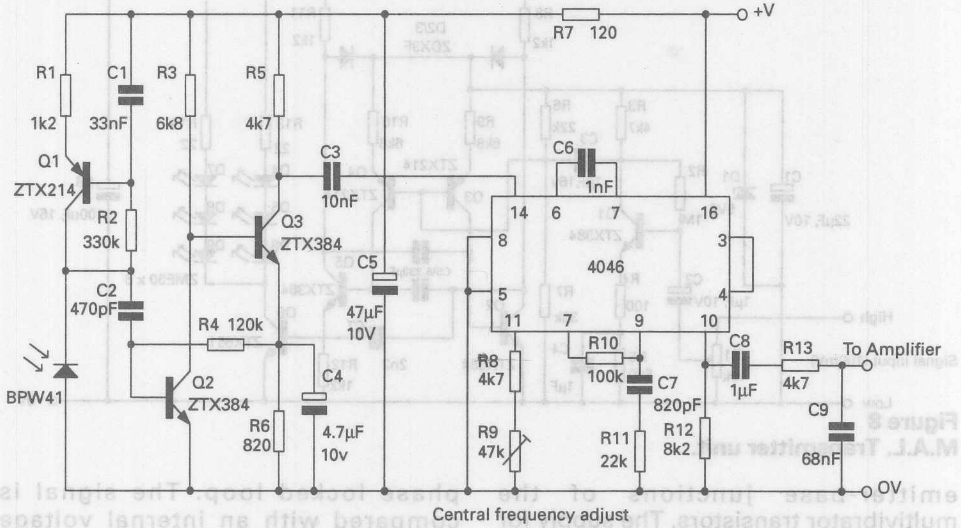


Figure 9a
M.A.L. Receiver and Detector.

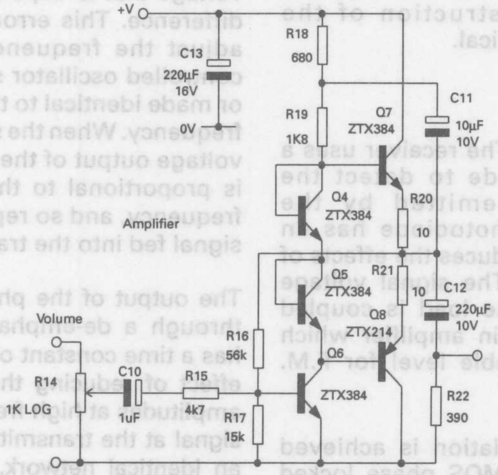


Figure 9b
M.A.L. Receiver Amplifier.

The corrected audio signal is then fed to an audio amplifier via a volume control.

The audio amplifier is intended to drive headphones with an impedance of 4Ω to 16Ω. It has a complementary push-pull output stage which employs a temperature compensated biasing circuit.

The receiver active load and amplifier circuits should be carefully laid out to minimise output to input feedback paths, otherwise instability may occur. The receiver should also be mounted in a metal case with a hole cut for light entry to the BPW41D.

M.A.L. Setting Up

The receiver phase locked loop centre frequency can be adjusted to the transmitter output frequency either by:

i) using an oscilloscope and comparing the transmitter output frequency to the frequency on the receiver pll device pin 4

or,

ii) by feeding a low level audio signal into the transmitter, gradually covering the photodiode with an opaque material and adjusting the centre frequency control for best signal recovery.

Once this adjustment has been made, the transmitter input level control should be set to give the maximum audio signal output before distortion due to overloading occurs.

Beam Interruption Detectors (B.I.D.)

Beam interruption detectors have a variety of uses ranging from object counting to burglar and fire detection alarms. They generally consist of a transmitter that emits a beam of light to a detector. Should this beam be broken, the detector gives a change in output level. The detector must discriminate between the transmitted beam and other light sources. This can be achieved by using a lensed system to make the detector directionally selective, or, by modulating the transmitter output and then making the detector select only correctly modulated sources. The lensed system uses simple transmitter and detector circuitry but mechanical construction and installation alignment are necessarily critical. The modulated system has the advantage of being easier to set up, but requires more complicated transmitter and detector circuitry.

A beam interruption detector using the BPW41D photodiode in a modulated system has been developed. It includes a transmitter circuit and a choice between two detector circuits of differing range capability. All the circuits are powered from a 5V d.c. supply to facilitate their operation with standard integrated circuit logic devices.

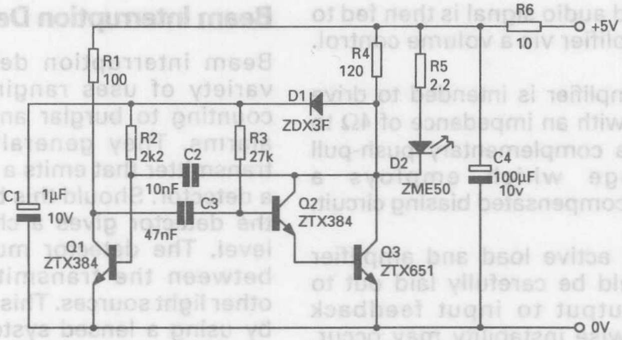


Figure 10
B.I.D. System transmitter.

B.I.D. Transmitter

The transmitter consists of an oscillator which drives a high output IR LED (ZME50). The oscillator is a sure-start multivibrator circuit which provides an output of 15 to 1000 mark-space ratio at a frequency of 1kHz. This large mark-space ratio allows the infra-red diode to be operated at a high peak current so as to maximise the range of the link. A decoupling network is included in the power supply of the transmitter to isolate it from any logic circuitry using the same 5V power source. The transmitter supply current is approximately 65mA. The circuit is illustrated in Figure 10.

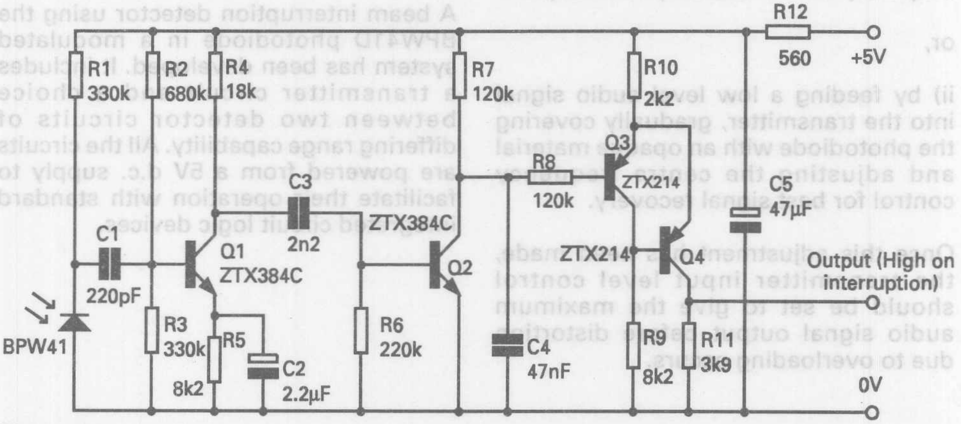


Figure 11
B.I.D. System receiver for short range duty.

Beam Interruption Short Range Detector

The short range detector circuit shown in Figure 11 is intended for use in parts counting and similar applications. A BPW41D photodiode with a simple resistor load is used to detect the transmitted beam. An active load is not used because it is usually easy to screen out interfering light sources in short range systems. The output of the photodiode is coupled to a single stage high input impedance amplifier which drives a pulse detector. The output of the pulse detector is low whilst a sufficiently strong transmitter beam is received, and high once this beam is interrupted.

The maximum range from the transmitter over which this detector will operate is approximately 0.7m. If necessary this can be increased by using a small lens with the transmitter L.E.D. to effectively increase its output. Using a 16mm dia. lens of 10mm focal length, the range using this detector was increased to 6m. It should be remembered however, that using a lens on the transmitter will make it's alignment more critical. The construction of this detector is not too critical, but electrical screening may be necessary in industrial environments.

Beam Interruption Long Range Detector.

This longer range detector can be used in applications such as burglar alarms, automatic door or gate controllers etc. In this circuit (shown in Figure 12) the BPW41D photodiode is used with an active load because ambient lighting conditions are much less controllable in long range systems, and this load will be unaffected by high background light levels. The improvement in range over the simple detector circuit is achieved by using a higher gain amplifier to boost the output of the BPW41D photodiode before feeding it to a pulse detector.

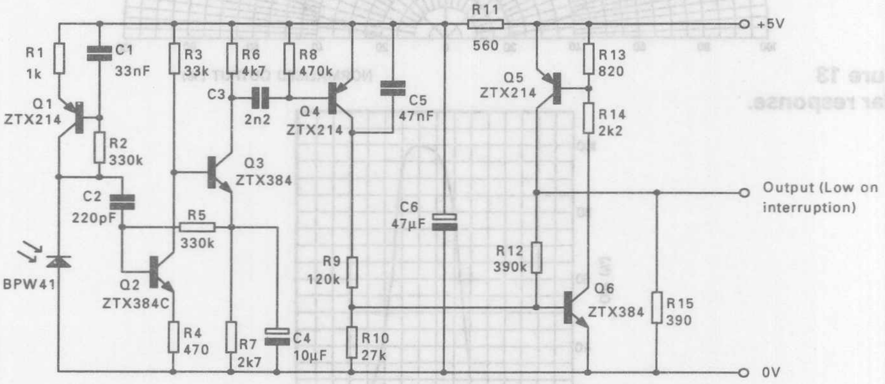


Figure 12
B.I.D. System receiver for long range duty.

The pulse detector drives a level sensor which produces a low output when the transmitter beam is broken. A small percentage of the output is fed back to the input of the level sensor to give the hysteresis necessary for noise free output switching. The level sensor output is capable of driving a normal TTL logic gate if required. The maximum range of operation with this detector is approximately 6m. Using a lens with the transmitter as described with the short range detector, a range of over 40m has been achieved.

During construction the active load and amplifier should be carefully laid out to avoid unwanted feedback. The unit should be mounted in a screened box to minimise interference.

The BPW41D is one member of the Zetex range of opto-electronic products, which includes a range of photo-diodes, photo-emitters and photo-transistors.

Appendix

Selected BPW41D datasheet charts.

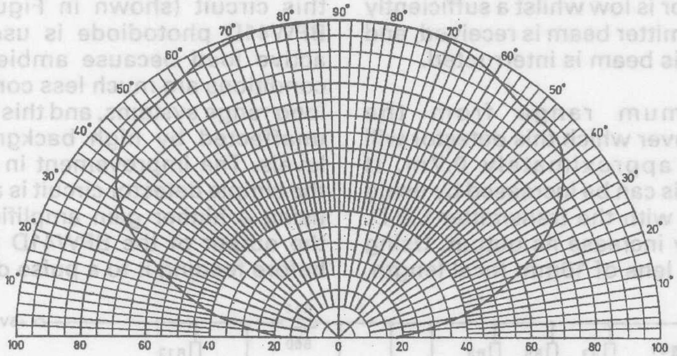


Figure 13
Polar response.

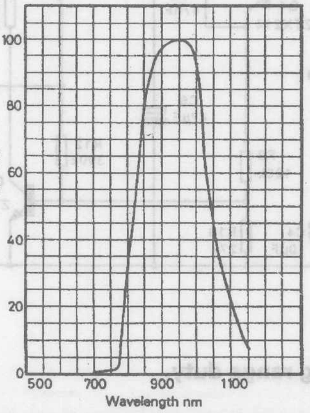


Figure 14
Relative Spectral Response.

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Application Note 4
Issue 2 January 1996

Centre Collector E-Line Applications

An Overview of Package Capability and Applications Advantages

David Bradbury

Introduction

This Application Note describes the product range, important package characteristics (including construction and thermal characteristics) and typical applications of the Zetex Semiconductors range of Centre Collector E-Line transistors.

Originally, the extensive range of ZETEX E-Line bipolar and MOSFET transistors was only available in the conventional edge collector (or drain) configuration. The high peak current, temperature and power capabilities exhibited by many of these devices allowed them to be used to replace higher power packages (TO126, TO202, TO220, TO237, SOT89, DPAK, etc.), which are mainly centre collector types. This fact, and second sourcing constraints, meant that circuit boards sometimes had to be laid out to accept both E-Line and larger centre collector power packages, wasting space and increasing costs. However, many of the high performance E-Line devices can now be supplied using centre collector/drain terminations, thus easing second sourcing problems for the designer.

Product Range

Design features of devices in the new Centre Collector E-Line range:

- NPN and PNP bipolar types up to 400V BV_{CEO}.
- NPN and PNP Darlingtons up to 160V BV_{CEO}.
- N and P channel MOSFET transistors up to 450V BV_{DSS}.
- High current gain.
- Super-β types available.
- High current handling capability - 10A peak, 3A continuous.
- Low saturation voltage.
- Fast switching and high F_T.
- T_j max up to 200°C.
- P_D up to 1W (this can be greater, and is very dependent on mounting method.)
- Small outline E-Line (TO92 style) package.
- Various through hole lead forms available as standard.
- Proven reliability and quality.

These features have been obtained by using advanced chip designs, the latest fabrication techniques and extensive modern production facilities, coupled with rigorous reliability and quality monitoring.

Further details of the product range can be found in the Zetex Semiconductor Data Book #1, Through Hole Components.

Package Construction and Reliability

The Centre Collector E-Line package is formed by transfer moulding a TO92 frame in a silicone plastic.

This plastic is specially selected to provide a rugged one-piece encapsulation resistant to severe environments, and allow the high junction temperatures normally associated with metal-can devices. The high glassing temperature and absence of ionic contaminants are two of the many features of the silicone encapsulant that allows chip operation up to 200°C without the risk of failure. The E-Line manufacturing process includes a resin backfill stage, involving vacuum impregnation of the moulded package with a silicone resin. This resin seals any voids which may exist between the lead frame and encapsulant, protecting the chip from exposure to corrosive agents and moisture, giving the package hermeticity properties approaching those of metal-can devices.

The resulting reliability given by these highly-developed packaging techniques and the use of well-passivated chips is routinely monitored. This monitoring covers tests such as high temperature reverse bias (HTRB), power cycling, temperature cycling and reliability in hazardous environments.

Power Rating

The Centre Collector TO92 lead frame was selected to allow a wide range of chip sizes and geometries, yet provide a low thermal resistance to the device. As can be expected, the thermal resistance offered by devices in the Centre Collector range depends on the chip size and, equally importantly, the mounting method. The centre collector range of devices was originally conceived to provide lead compatibility to allow drop in replacement of TO220 and TO126 packaged components, when these parts were being used for their high current capability rather than in a linear mode, (e.g. power dissipative). The thermal resistance presented by the E-Line will be higher than for a much larger packaged component, but this is often compensated for in most practical circumstances by the E-Line package's capability to be used with junction temperatures up to 200°C.

The thermal characteristics of the Centre Collector E-Line package is shown in Figure 1.

Figure 1 shows the power derating curves for an FXT651. For comparison purposes, the derating curves for free standing TO92, TO237 and TO220 types are also included in this diagram. Note that the E-Line package can provide excess power capability over TO92 and TO237 types at any temperature, and even over TO220 types at temperatures that occur in many "real" applications such as within automotive environments.

Full DC and transient thermal resistance data are available for all devices in the Centre Collector E-Line range.

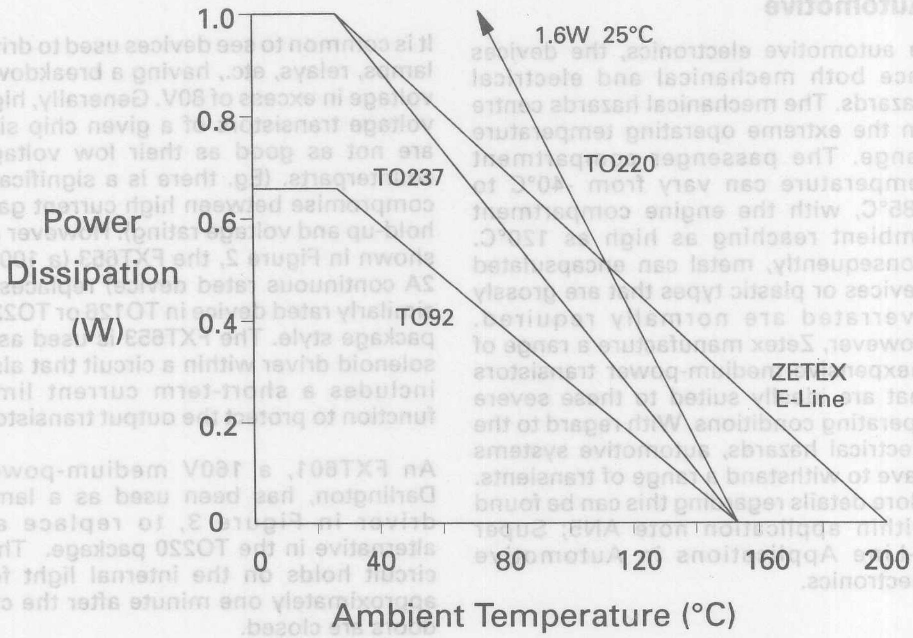

Figure 1
Power Derating Curves for Through Hole Mounted Free Standing Devices.

Applications

Some application areas that can take advantage of the E-Line Centre Collector package are listed below:

Automotive - Lamp and display drivers, Solenoid and relay drivers, Motor drivers.

Converters - Capacitor charging (Flash guns, Ignition units), Logic supplies, Isolated supplies, DC-AC supplies (Fluorescent lamps).

Computer Peripherals - Display and lamp drivers, Stepper Motor drivers (printers), Solenoid and relay drivers.

Telephones - Pulse code diallers, Earpiece muting, Hookswitches.

Miscellaneous - Low-level audio output stages, Low power RF transmitters, Frame drivers, Buzzers, Radio control, Warning lamps, Gaming machines.

Provided overleaf are some typical application examples mentioned in the above list.

Automotive

In automotive electronics, the devices face both mechanical and electrical hazards. The mechanical hazards centre on the extreme operating temperature range. The passenger compartment temperature can vary from -40°C to +85°C, with the engine compartment ambient reaching as high as 120°C. Consequently, metal can encapsulated devices or plastic types that are grossly overrated are normally required. However, Zetex manufacture a range of inexpensive medium-power transistors that are ideally suited to these severe operating conditions. With regard to the electrical hazards, automotive systems have to withstand a range of transients. More details regarding this can be found within application note AN5; Super E-Line Applications in Automotive Electronics.

It is common to see devices used to drive lamps, relays, etc., having a breakdown voltage in excess of 80V. Generally, high voltage transistors of a given chip size are not as good as their low voltage counterparts. (Eg. there is a significant compromise between high current gain hold-up and voltage rating). However as shown in Figure 2, the FXT653 (a 100V, 2A continuous rated device) replaces a similarly rated device in TO126 or TO220 package style. The FXT653 is used as a solenoid driver within a circuit that also includes a short-term current limit function to protect the output transistor.

An FXT601, a 160V medium-power Darlington, has been used as a lamp driver in Figure 3, to replace an alternative in the TO220 package. This circuit holds on the internal light for approximately one minute after the car doors are closed.

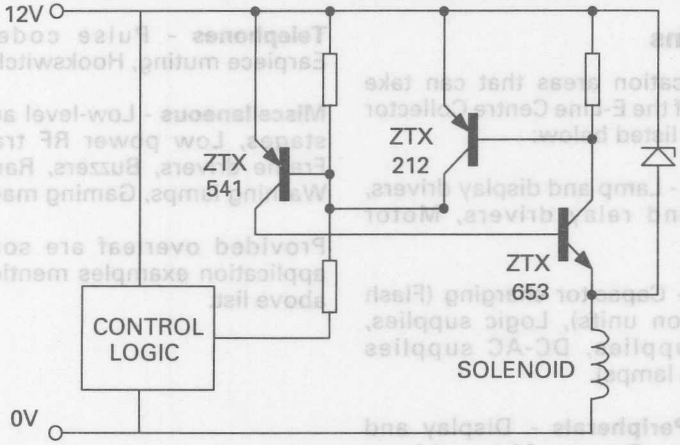

Figure 2
Automotive Systems Relay Driver using Centre Collector E-Line.

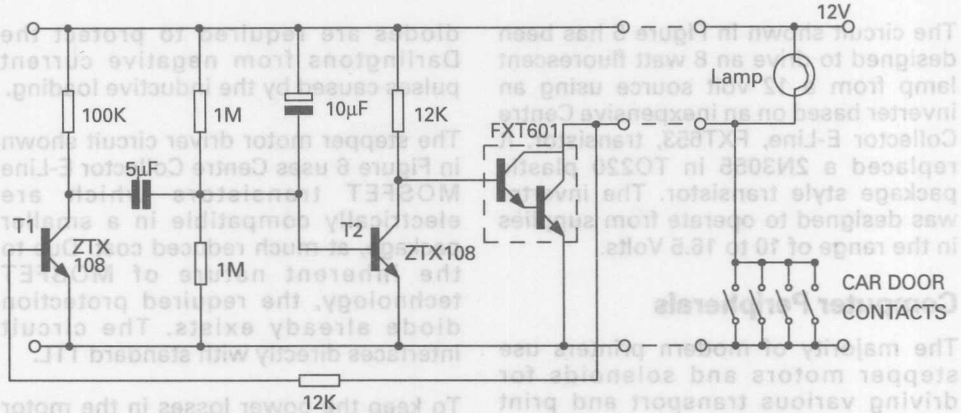

Figure 3
Automotive System Courtesy Lamp Driver using a Centre Collector Darlington Transistor.

Converters

Figure 4 shows a circuit diagram for a portable, low-power beacon used for distress signalling systems. A 3 Volt D.C. supply is converted to 350 Volt D.C.

The original unit employed a TO220 package device which gave a poor conversion efficiency. The superior characteristics of the Centre Collector E-Line transistor, FXT653, allowed the unit to be operated at a higher frequency, thus reducing the size of the magnetics. The end result was a circuit with high efficiency, smaller size and a significant cost saving.

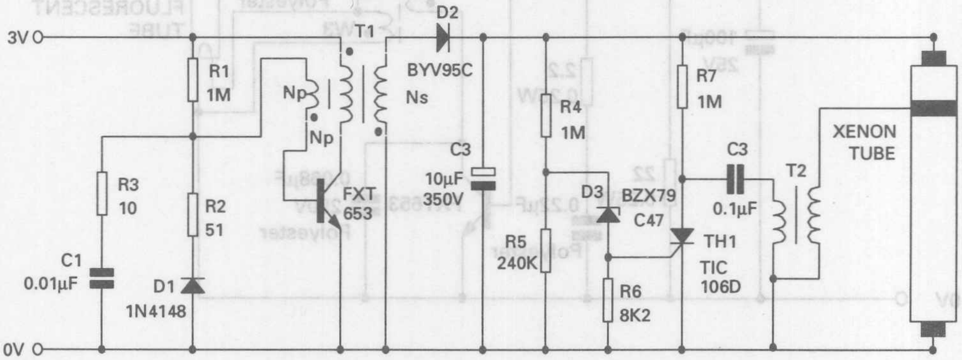

Figure 4
Low Power Xenon Beacon employing an FXT653 in a 3V DC to 350V DC Converter.

The circuit shown in Figure 5 has been designed to drive an 8 watt fluorescent lamp from a 12 volt source using an inverter based on an inexpensive Centre Collector E-Line, FXT653, transistor. It replaced a 2N3055 in TO220 plastic package style transistor. The inverter was designed to operate from supplies in the range of 10 to 16.5 Volts.

Computer Peripherals

The majority of modern printers use stepper motors and solenoids for driving various transport and print mechanisms. For ease of driving and circuit simplicity, expensively packaged Darlingtons with integral collector-emitter diodes are used. The diodes are required to protect the Darlingtons from negative current pulses caused by the inductive loading.

The stepper motor driver circuit shown in Figure 6 uses Centre Collector E-Line MOSFET transistors which are electrically compatible in a smaller package, at much reduced cost. Due to the inherent nature of MOSFET technology, the required protection diode already exists. The circuit interfaces directly with standard TTL.

To keep the power losses in the motor low during periods of low speed operation, a logic controlled positive line switch is required. The FXT704 PNP Darlington transistor can easily be used

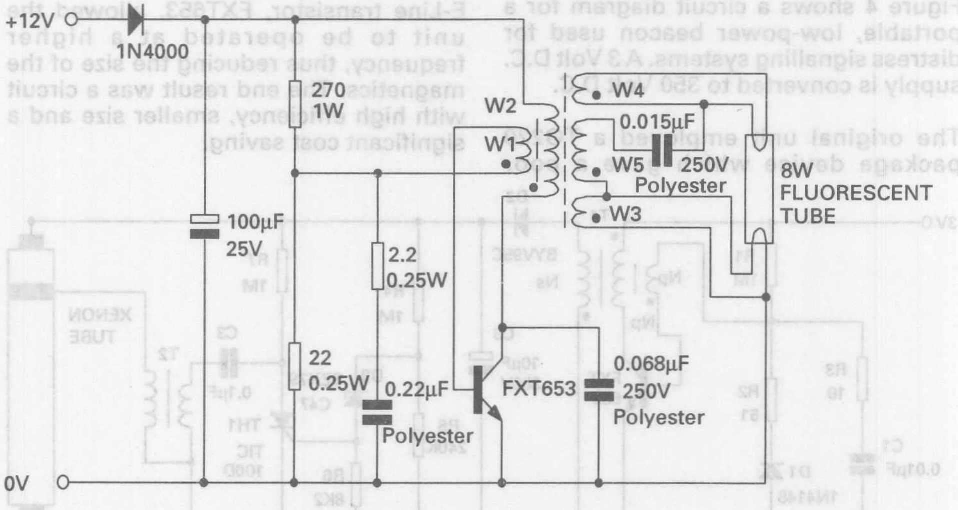

Figure 5
8W Fluorescent Lamp Inverter.

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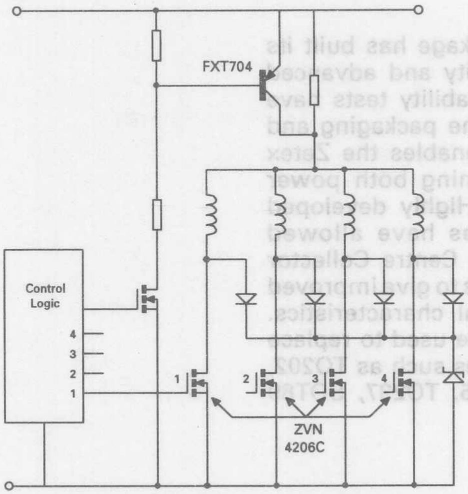


Figure 6
Centre Drain MOSFETs utilised within a Stepper-Motor Module.

in this application for motor currents to 4A peak. The motor circuit shown interfaces directly with standard TTL.

Figure 7 shows the typical voltage and current waveform of a MOSFET used in the motor drive.

Telephones

High breakdown voltage, high current handling capability, high gate input impedance, low on-resistance and low cost make Centre Collector E-Line MOSFETs an ideal choice for use as dial-pulse and muting switches in a dialler circuit. Figure 8 shows the use of high voltage Centre Collector E-Line bipolar and MOSFETs in a typical telephone instrument.

The Centre Collector E-Line package used in this application replaces commonly-used TO237 plastic package.

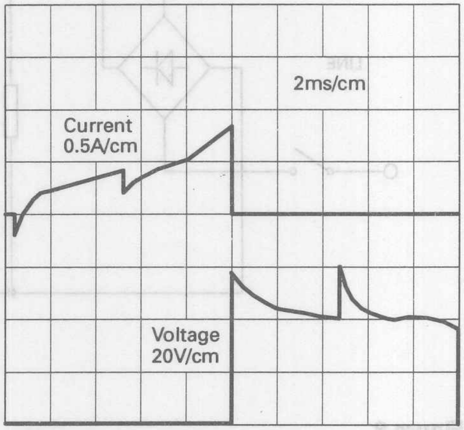


Figure 7
Stepper Motor Drive Circuit - Source Current and Drain-source Voltage of ZVN4206C During Drive Cycle.

Summary

The Zetex E-Line package has built its reputation on reliability and advanced design. Extensive reliability tests have been performed on the packaging and die geometries that enables the Zetex E-Line range, confirming both power ratings and quality. Highly developed packaging techniques have allowed Zetex to produce the Centre Collector range of E-Line devices to give improved thermal and electrical characteristics. This allows them to be used to replace higher power packages such as TO202, TO220, TO126, TO225, TO237, SOT89 and DPAK.

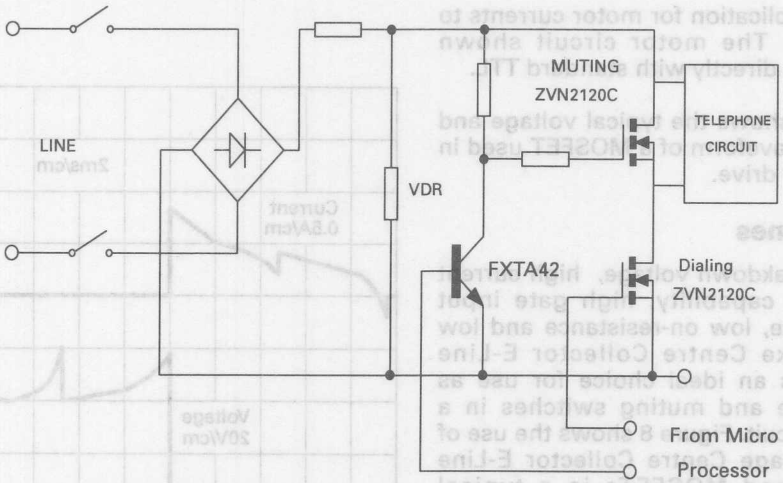


Figure 8
Centre Collector High Voltage MOSFETs and Bipolar transistors within a Telephone.

Application Note 5
Issue 2 February 1996

Super E-Line Applications in Automotive Electronics

Replacement of Large Packaged Transistors with an Enhanced TO92 Product

David Bradbury

Car buyers are now demanding greater and greater sophistication in their purchases, which along with extra safety features, pollution controls, security systems and manufacturing cost-cutting measures, are leading to an ever-increasing use of electronics in automobiles. Automobile electronics have extended from the occasional radio, to cover every major area within a vehicle. Engine management, fuel injection system controls, transmission controls, anti-lock braking systems, electronic displays, warning systems, etc., are just a few examples of the wide range of electronic systems that can be presently found in modern cars.

Many of these systems involve high current loads such as relays, solenoids, lamps, displays, motors and audible warning systems. These require one or more power transistors to interface between the controlling logic and the load. This brief note explains some of the major problem areas in selecting devices and circuits for automobile use, and demonstrates the use of the Zetex Super E-Line range of transistors in some practical examples.

Problem Areas

The hazards for power devices in automotive electronic systems are both mechanical and electrical. The mechanical hazards centre on the extreme operating temperature range endured by the devices.

To help designers evaluate the performance of Super E-Line transistors as temperatures vary, Zetex can supply a wide range of temperature-related data on both standard and special characteristics. Figure 1 shows the standard data sheet $V_{CE(sat)}$ against I_C graph for a ZTX602 Super E-Line Darlington transistor. Note how the characteristics are shown for a range of temperatures. Even when this data is not given on the data sheet, Zetex can supply this quickly for customers who require such information.

Mechanical Requirements

Passenger compartment temperatures can vary from - 40°C to + 85°C with the engine compartment ambient reaching as high as 120°C. High ambient temperatures place serious constraints on the performance and reliability of any power semiconductors used. As a result, Metal Can encapsulated devices or plastic types that are grossly overrated are normally required. However, Zetex manufacture a range of inexpensive medium power transistors that are ideally suited to these severe operating conditions.

Zetex Super E-Line and E-Line bipolar transistors are encapsulated in a special Silicone plastic that allows operation at junction temperatures up to 200°C. This gives them the temperature performance of Metal Can devices with the robustness and low cost of a plastic encapsulation. Figure 2 illustrates some of the advantages given by the E-Line package. It compares the power rating against ambient temperature of the Super E-Line 1W package, with standard TO92, TO237 and TO220 style packages.

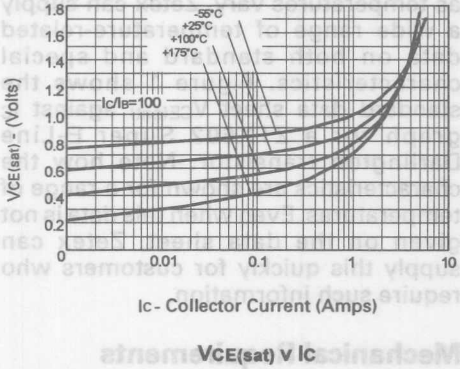


Figure 1
Typical Collector Emitter Saturation Voltages vs Collector Current, with Temperature as a Parameter.

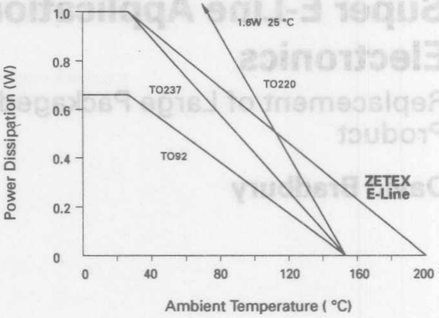


Figure 2
Power Derating Curves Showing Power Dissipation Advantage of the E-Line encapsulation.

Note how standard TO92 and TO237 types give inferior performance to that of the E-Line range, with even the TO220 style devices requiring a heatsink to match E-Line types in ambient temperatures above 110°C. Also, even higher levels of power dissipation are permissible if the E-Line devices are mounted on circuit boards with a low thermal resistance. For instance, devices from the Super E-Line range can dissipate 1.5W when mounted on a circuit board providing 1 square inch copper area to the collector termination. Furthermore, pulse applications allow higher powers still. Figure 3 shows the actual power dissipation permissible for a ZTX650 mounted on a 1 inch square board for a range of pulse widths. This chart can be used for design analysis when using any of the ZTX649-658 and ZTX749-758 transistor ranges.

Extensive reliability testing has been done on the packaging and chips that make up the Zetex E-Line range, confirming both power ratings and quality. Temperature cycling, climatic and electrical life evaluation tests are routinely carried out on all of the Zetex discrete products to guarantee continuing quality.

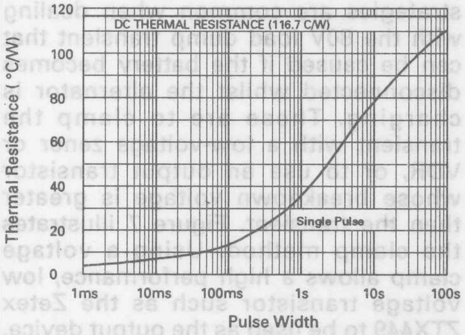


Figure 3
Transient Thermal Resistance for a ZTX650 Series Transistor Mounted on a 1" square FR4 PCB.

Electrical Requirements

Before moving on to some application examples, it is worth highlighting some of the electrical hazards that affect their design. Apart from normal 10-15V operation, automotive electronic systems have to withstand a wide range of transients. These include reverse battery connection, double voltage supplies, a 240V/0.5J positive line transient, an 80V/50J load dump transient, a -100V/0.5J supply disconnect transient and many other lower energy but higher voltage transients. Figures 4 and 5 show the voltage waveforms of the first two of these transients. A further requirement is that the system can withstand a reasonable amount of mistreatment during maintenance. Accidents do happen and efforts should be made to limit their seriousness.

Application Examples

The design of load driver circuits which operate using a vehicle's normal battery voltage range, comes from a fairly straightforward consideration of drive levels, output current and power dissipation.

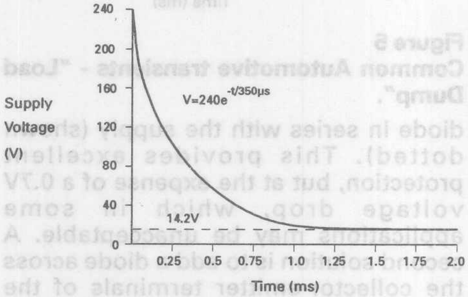


Figure 4
Common Automotive Transients - 240V Line Transient.

Reverse battery protection and transient protection, however, can be achieved in number of ways, with occasionally some conflict of effect. In the following application examples, consideration is given to the problems of designing load driver circuits for automobile use.

Figure 6 shows a simple unprotected lamp driver circuit. In this circuit, a reverse connected battery would force current through the collector-base junction of the output transistors and into the control IC, with possibly disastrous consequences. One good solution to this problem is to insert a

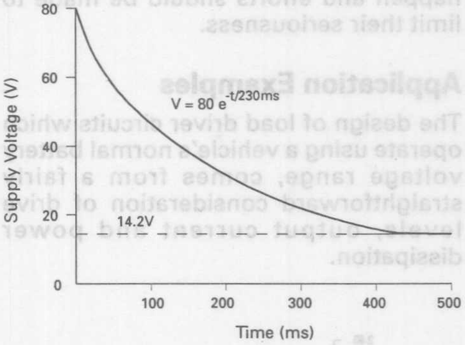


Figure 5
Common Automotive transients - "Load Dump".

diode in series with the supply (shown dotted). This provides excellent protection, but at the expense of a 0.7V voltage drop, which in some applications may be unacceptable. A second solution is to add a diode across the collector-emitter terminals of the output device. This diode shorts the fault current to ground, thus protecting the transistor and drive IC without causing any voltage drop in normal operation. However, which reverse battery protection technique used may well depend on the type of transient protection employed.

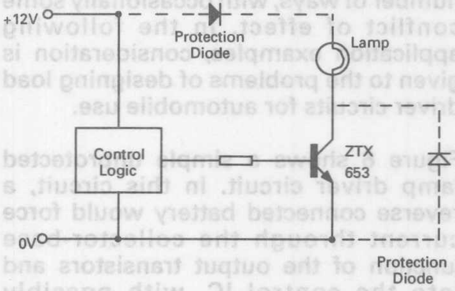


Figure 6
Reverse Battery Protection.

The wide range of transients that occur within an automobile often require a number of different protection techniques to be employed. Two strategies are common when dealing with the 80V load dump transient that can be caused if the battery becomes disconnected whilst the alternator is charging. These are to clamp the transient with a low-voltage zener or VDR, or to use an output transistor whose breakdown voltage is greater than the transient. Figure 7 illustrates the clamp method. Using a voltage clamp allows a high performance, low voltage transistor such as the Zetex ZTX449 to be used as the output device. The clamp component has to withstand a current that is several times greater than the normal current taken by the load. This can cause very high power dissipation in the clamp for heavy loads, so the technique is normally restricted to lower current load drivers.

Using a high voltage transistor (i.e. greater than 80V) makes a clamp

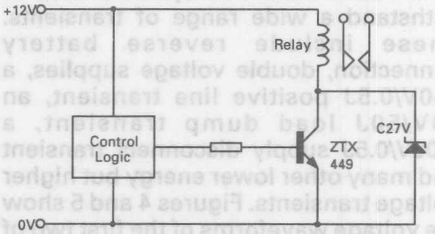


Figure 7
Transient Clamp using a Low Voltage Driver and High Current Zener Diode.

unnecessary for the load dump transient. The current handling capability of high voltage transistors of a given chip size are generally not as good as their low voltage counterparts, but Zetex can offer suitable Super E-Line transistors with a 2A continuous rating (The 100V ZTX653 for instance). Using a high voltage transistor does not completely eliminate the need for some form of clamp circuit. Transients as high as 500V can occur in an automobile, and measures must be taken to limit these. However, these high voltage transients are low energy types which can be easily controlled using a low power clamp circuit. Thus, the circuit when using a high voltage output transistor (shown in Figure 8) looks identical to the low voltage clamp circuit of Figure 7, however the power rating of the clamp component (and it's cost) will be much lower.

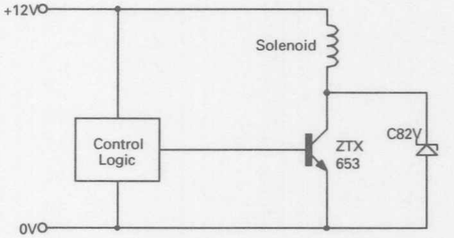


Figure 8
High Voltage Driver with Low Current Rated Transient Clamp.

Any load that can be removed from the circuit, such as a lamp bulb that can be unscrewed or a relay that can be unplugged, may be a potential reliability hazard during maintenance. Accidental shorts can all too easily happen (cleaning connectors with a screwdriver, for instance) and so it can be very worthwhile to protect the driver circuit against this possibility. Figure 9 shows an output circuit which includes a short-term current limit function to protect the output transistor.

The characteristics of the circuit block labelled CONTROL LOGIC in the previous diagrams can be important when selecting output devices. Often this element is constructed using low output current MOS or CMOS circuits. Since many loads such as lamps and relays can under some circumstances take surge currents in excess of 1 A, the output devices used must either employ interface circuits or have very high current gain. This high gain can be achieved by using MOSFET or Darlington transistors. For instance, the Zetex ZVN4206 MOSFET transistor could be substituted for the ZTX449 in Figure 7, allowing a low output current logic circuit to be used. Zetex also supply NPN and PNP Darlingtons that are

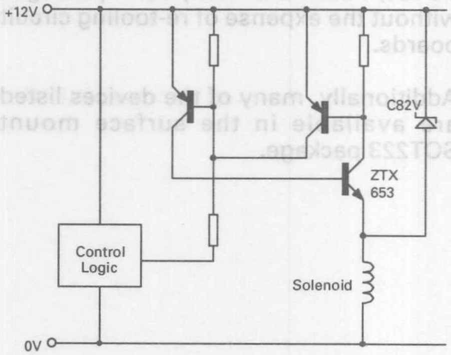


Figure 9
Load driver with current limit.

equally suitable for these applications. Figure 10 shows a four-phase stepping motor driver using ZTX600 Darlington transistors, directly driven from a CMOS microprocessor.

ZETEX Super E-Line Product Range

Tables 1 and 2 within Appendix A show some of the wide range of Zetex Super E-Line devices that can be useful in automotive applications. The tables can be used to choose the best device to suit the particular circuit requirements of voltage, current and gain. The gain and current specifications of a load driver are normally easy to calculate, but as mentioned previously, the breakdown voltage required is dependent on the transient protection method used.

Most medium power E-Line and Super E-Line devices can now also be supplied in the centre collector lead configuration. This allows the superior performance of the Zetex Super E-Line and E-Line range to be utilised in replacing TO202, TO220, TO237, SOT89, TO126, TO225 and many other packages without the expense of re-tooling circuit boards.

Additionally, many of the devices listed are available in the surface mount SOT223 package.

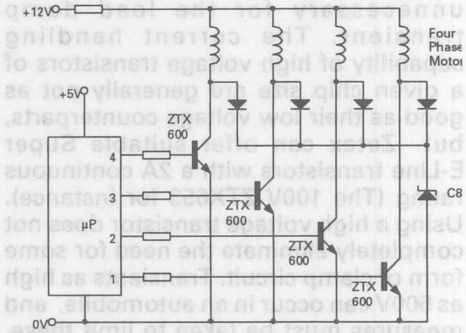


Figure 10
Four Phase Stepper Motor Drive Circuit using ZTX600 Darlington Transistors.

Appendix A

A selection of some of the Zetex range of products particularly suited to Automotive applications. For full characterisation refer to Data Book.

NPN PNP	ZTX649 ZTX749	ZTX650 ZTX750	ZTX651 ZTX751	ZTX652 ZTX752	ZTX653 ZTX753	UNITS
V_{CBO}	35	60	80	100	120	V
V_{CEO}	25	45	60	80	100	V
V_{EBO}	5					V
I_{CM}	6					A
I_C	2					A
P_{TOT}	1					W
h_{FE} min	75 @ I_C=2A, V_{CE}=2V	80 @ I_C=1A, V_{CE}=2V		55 @ I_C=1A,V_{CE}=2V		
h_{FE} min	15 @ IC=6A, V_{CE}=2V	40 @ I_C=2A, V_{CE}=2V		25 @ I_C=2A, V_{CE}=2V		
$V_{CE(sat)}$ max	0.5V @ I_C=2A, I_B=0.2A					V

Table 1
ZTX649-653 & ZTX749-753 Series Main Parameters.

NPN PNP	ZTX602	ZTX603	ZTX604 ZTX704	ZTX605 ZTX705	UNITS
V_{CBO}	80	100	120	140	V
V_{CEO}	60	80	100	120	V
V_{EBO}	10				V
I_{CM}	4				A
I_C	1				A
P_{TOT}	1				W
h_{FE} min	500 @ I_C=2A, V_{CE}=5V		2000 @ I_C=1A, V_{CE}=5V		
h_{FE} min	2000 @ I_C=1A, V_{CE}=5V				
$V_{CE(sat)}$ max	1.0V @ I_C=1A, I_B=1mA		1.5V @ I_C=1A, I_B=1mA		V

Table 2
ZTX602-605 & ZTX704-705 Series Main Parameters.

Notes

The Use of Zetex E-Line Transistors in Motor and Solenoid Functions within Printers

Effective Logic to High Current Load Interfacing

Neil Chadderton

Introduction

The majority of modern printers use stepper motors and solenoids for driving the various transport and print mechanisms. These applications usually involve using drive signals directly from the control logic; and handling high-pulsed currents and voltages. In the past this has led to TO220 and similar devices being used, but an alternative is now available. Devices from the Super E-Line range of transistors are direct replacements for the aforementioned devices in medium power situations. This Application Note outlines a few examples that are applicable and is by no means comprehensive.

Motor Drivers

Stepper motors are used for many printer functions, such as platten roller and print head position drivers, and daisy wheel drivers.

A typical circuit used in many printers is shown conceptually in Figure 1. The driver transistors Q1-Q4 are switched in an ordered sequence to energise the motor's coils to produce the desired action. As the transistor load is inductive, the voltage present on switching off easily exceeds the supply voltage, while the current flowing in the on-state follows a ramp-like appearance.

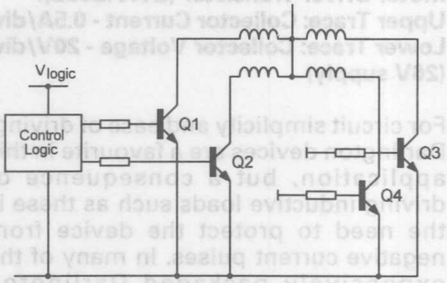


Figure 1
Typical Stepper Motor Drive Circuit (diagramatic).

These conditions are illustrated in Figure 2, which shows the voltage and current experienced by the transistor in a typical cycle. (In this case, a centre drain MOSFET, the ZVN4206C has been employed).

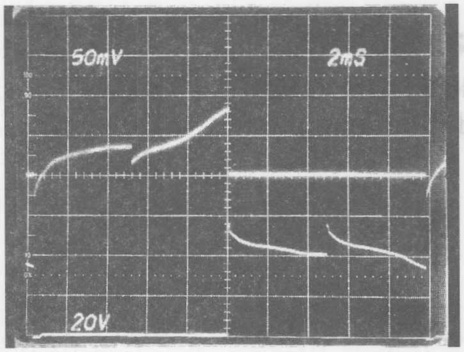


Figure 2
Typical Waveforms on a 4-Phase Stepper Motor Driver Transistor (ZVN4206C).
Upper Trace: Collector Current - 0.5A/div.
Lower Trace: Collector Voltage - 20V/div.
(26V supply)

For circuit simplicity and ease of driving, Darlington devices are a favourite in this application, but a consequence of driving inductive loads such as these is the need to protect the device from negative current pulses. In many of the expensively packaged Darlington transistors, this is effected by an integral collector-emitter diode.

An alternative is to use MOSFET transistors, which exhibit all the electrical requirements in a smaller package and at much reduced cost, and possess the necessary protection diode due to the inherent nature of MOSFET technology.

Even in very demanding driving applications for larger printer motors such as illustrated in Figure 3, E-Line transistors can be used to obtain a large cost advantage.

The negative current pulses due to the driving configuration, and the high current involved (greater than 3A), make this an ideal situation for a high performance E-Line Darlington transistor. Used with an inexpensive externally-wired diode, this offers up to a four-times cost-saving on the expensive TO220 alternative.

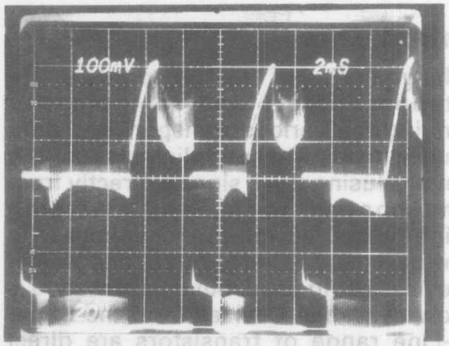


Figure 3
Waveforms on Stepper Motor Driver Transistor for a Large Printer (FXT605).
Upper Trace: Collector Current - 1A/div.
Lower Trace: 20V/div.

Dot Matrix Head Drivers

For the dot matrix form of printer, solenoids are usually used to force a pin to strike the ribbon and thereby produce a dot on the paper. A column of these pins is driven as required to match the matrix representation of printable characters.

As in the motor driver example, the solenoid also presents an inductive load and Figure 4 typifies the circuit used to drive it. A control pulse from the logic turns the transistor on to energise the solenoid coil.

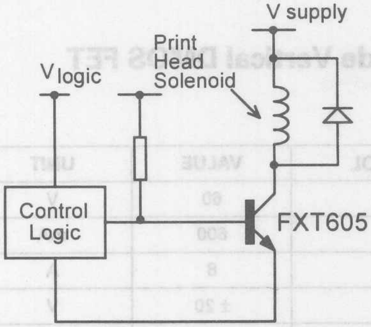


Figure 4
Typical Circuit of a Dot Matrix Head Driver.

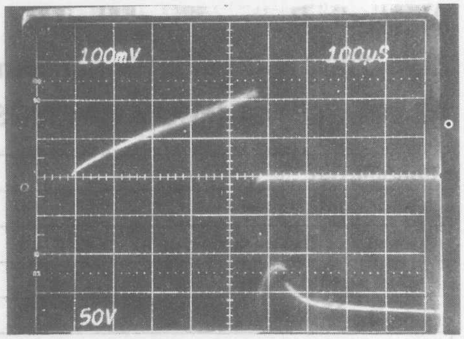


Figure 5
Waveforms on a Dot Matrix Head Driver Transistor (FXT605).
Upper trace: 1A/div. Lower trace: 50V/div.

The waveforms shown in Figure 5 were obtained for a dot matrix head driver. The back-emf and the high current peaks demand a device of superior performance easily met by the Zetex range of E-Line devices.

Centre Collector E-line Transistors

Zetex Semiconductors offer a comprehensive range of centre collector E-Line transistors that can directly replace TO220 and similar devices in many applications. These devices match all the electrical requirements, and also show a large saving in size and cost.

Summary

Zetex transistors can be used with confidence in many motor and solenoid driver applications of which the above are examples. The state of the art transistor technology coupled with the excellent thermal characteristics of the E-Line package, and the convenience of the centre collector option, produce an attractive solution to an otherwise costly situation.

Appendix A

FXT605 Centre Collector NPN Power Darlington Transistor

SYMBOL	MIN.	MAX.	UNIT	CONDITIONS.
I_C I_{CM}		1 4	A A	Continuous Pulsed
$V_{(BR)CEO}$	120		V	I_C=10mA, I_B=0
$V_{CE(sat)}$		1.0 1.5	V V	I_C=0.25A, I_B=0.25mA I_C=1A, I_B=1mA
h_{FE}	2000 500			I_C=50mA, V_{CE}=5V I_C=2A, V_{CE}=5V
f_T	150		MHz	I_C=100mA, V_{CE}=10V f=20MHz

Appendix B

ZVN4206C N-Channel Enhancement Mode Vertical DMOS FET

ABSOLUTE MAXIMUM RATINGS

PARAMETER	SYMBOL	VALUE	UNIT
Drain-Source Voltage	V_{DS}	60	V
Continuous Drain Current at T_{amb}=25°C	I_D	600	mA
Pulsed Drain Current	I_{DM}	8	A
Gate-Source Voltage	V_{GS}	± 20	V
Power Dissipation at T_{amb}=25°C	P_{tot}	0.7	W
Operating and Storage Temperature Range	T_j:T_{stg}	-55 to +150	°C

ELECTRICAL CHARACTERISTICS (at Tamb = 25°C unless otherwise stated).

PARAMETER	SYMBOL	MIN.	MAX.	UNIT	CONDITIONS.
Drain-Source Breakdown Voltage	BV_{DSS}	60		V	I_D=1mA, V_{GS}=0V
Gate-Source Threshold Voltage	$V_{GS(th)}$	1.3	3	V	I_D=1mA, V_{DS}= V_{GS}
Gate-Body Leakage	I_{GSS}		100	nA	V_{GS}=± 20V, V_{DS}=0V
Zero Gate Voltage Drain Current	I_{DSS}		10 100	µA µA	V_{DS}=60V, V_{GS}=0V V_{DS}=48V, V_{GS}=0V, T=125°C(2)
On-State Drain Current(1)	$I_{D(on)}$	3		A	V_{DS}=25V, V_{GS}=10V
Static Drain-Source On-State Resistance (1)	$R_{DS(on)}$		1 1.5	Ω Ω	V_{GS}=10V,I_D=1.5A V_{GS}=5V,I_D=500mA
Forward Transconductance(1)(2)	g_{fs}	300		mS	V_{DS}=25V,I_D=1.5A
Input Capacitance (2)	C_{iss}		100	pF	V_{DS}=25V, V_{GS}=0V, f=1MHz
Common Source Output Capacitance (2)	C_{oss}		60	pF	
Reverse Transfer Capacitance (2)	C_{rss}		20	pF	

Both Appendix A and B are extracts from the *Discrete Through Hole Components Data Book* which detail full characterisation of the products quoted.

ZETEX

Application Note 7
Issue 2 January 1996

The Use of Zetex E-Line Transistors in DC-DC Converters

An Introduction and Typical Applications

David Bradbury

What Are DC-DC Converters?

DC-DC Converters are circuits designed to match loads to the power sources available. Generally, they accept power input at one voltage and supply one or more loads requiring power at different voltage levels. This description of converters is necessarily vague because of the extremely wide range of applications these circuits are used in.

Where Are They Used?

- In providing additional power supplies for applications such as computers, e.g. displays, memories, programmers, interfaces, etc.

- In matching power supplies to "difficult" loads such as capacitor charging in flash guns, pulsed laser drivers, ignition systems (spark generation), etc.

- In high efficiency voltage step down conversion.

- In safe high voltage power supplies.

Circuit Considerations and Device Characteristics

All the applications listed rely on the technique of using a switching circuit to turn the DC input into an AC signal, that can be conditioned using a transformer and rectifier to finally give the voltage and source characteristics required by the load. To obtain high efficiency, it is crucial that the transistor switch used to generate the AC signal gives very low losses. For low losses, the transistors must exhibit low conduction voltages and fast transition times. These must be achieved despite the high peak currents passed in most of the applications.

Size constraints are also very important. Unfortunately, small package size and high current handling are not characteristics that normally go together. A problem sometimes allied to size is operating temperatures. Circuits squeezed down to the limit may need to run at high temperatures, so this must be kept in mind when selecting a semiconductor switch.

Zetex offer a range of both bipolar and MOSFET transistors in the high reliability E-Line package with specifications that ideally suit many of the application areas listed previously.

This includes the Super E-Line range of NPN and PNP bipolar transistors with continuous current ratings up to 2A (6A peak) for 100V transistors and up to 0.5A continuous (1A peak) for a 300V device. Typical $V_{CE(sat)}$ voltages measure out at less than 250mV, even at these high continuous current levels. (**Important**: Please refer to **Note 1** on page AN7-5)

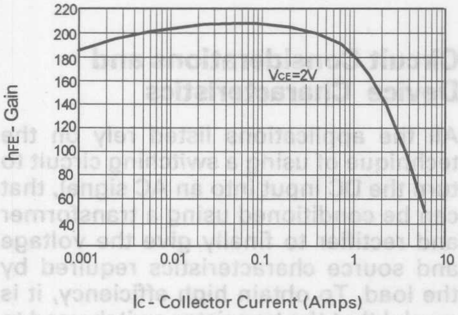


Figure 1
ZTX649 h_{FE} profile illustrating excellent gain holdup above 2A.

As an example, Figure 1 shows how the useful gain range of the 25V ZTX649 extends well beyond its 2A continuous rating. The bipolar E-Line 200°C maximum temperature rating and small package outline allows new converter designs to be more compact. Centre collector versions of the E-Line range can replace TO220 and SOT89 types presently used in many applications.

For circuits where very high speeds or ease of driving is important, Zetex also offer a useful range of MOSFET transistors encapsulated in the E-Line package. N-Channel types available presently are rated from 60V through 450V, and continuous current ratings up to 1.1A. This voltage range is duplicated for P-Channel types though their resistance is higher. The output characteristics of the 60V ZVN4206A are shown in Figure 2 to illustrate the high current capabilities of the Zetex E-Line MOSFET range. All these devices share the small outline and proven reliability of the Zetex bipolar E-Line range, however chip constraints presently limit the maximum junction temperatures of the MOSFET devices to 150°C.

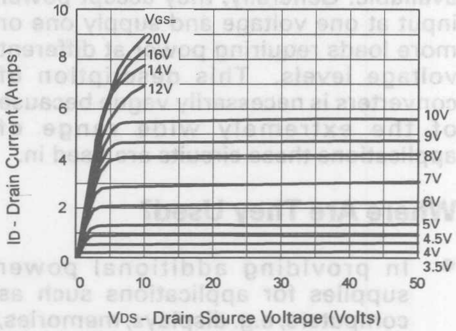


Figure 2
ZVN4206A Output Characteristics.

Using Zetex E-Line devices, DC-DC converters up to 10W output can be easily made using a single switching transistor, even when supplying "difficult" loads. More typical loads will allow higher outputs to be achieved. Figure 3 shows the circuit of an 8W fluorescent lamp inverter (strictly a DC-AC converter). A fluorescent lamp is a particularly awkward load for a converter since it has a highly non-linear

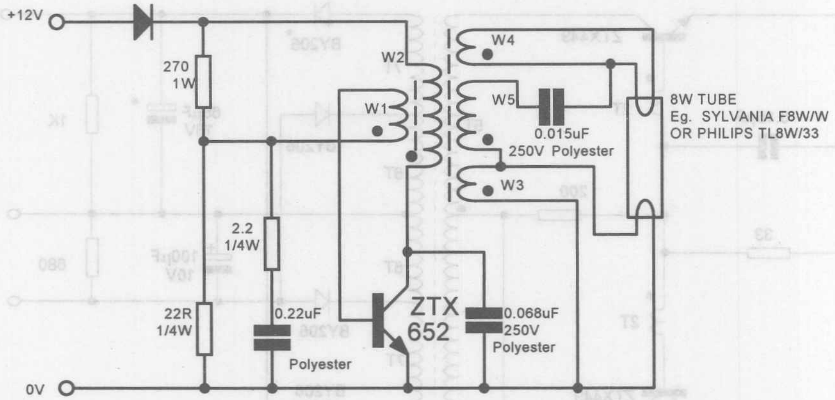


Figure 3
8W Output from a Single Switching Transistor.

impedance that varies greatly as the tube strikes and enters its normal operating condition. The converter must have a high output voltage capability for striking, yet also be capable of supplying the low impedance load the tube presents after it has struck. This leads to the specification of a high voltage transistor that can also pass high peak currents. These conflicting requirements lead many manufacturers to use a relatively expensive TO220 packaged device, but this application can be well served using a single Zetex E-Line ZTX652 transistor. Figure 4 shows the circuit waveforms obtained when using the ZTX652. Note how peak current levels approach 3A without high $V_{CE(sat)}$ voltages.

Figure 5 shows a very simple 5W converter intended to provide the +12V and -5V supplies often required in personal computers for memory, interface and display circuitry. Its power source is the standard 5V logic supply. The low value of this input supply means that relatively high peak currents for a given output power and low values of $V_{CE(sat)}$ for good regulation are required. In this forward converter operating at

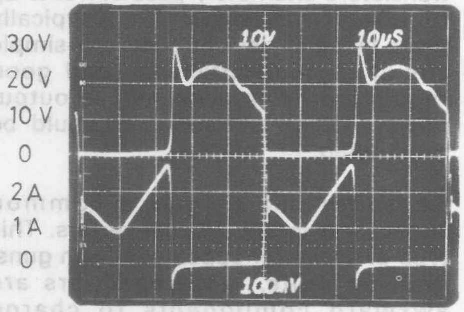
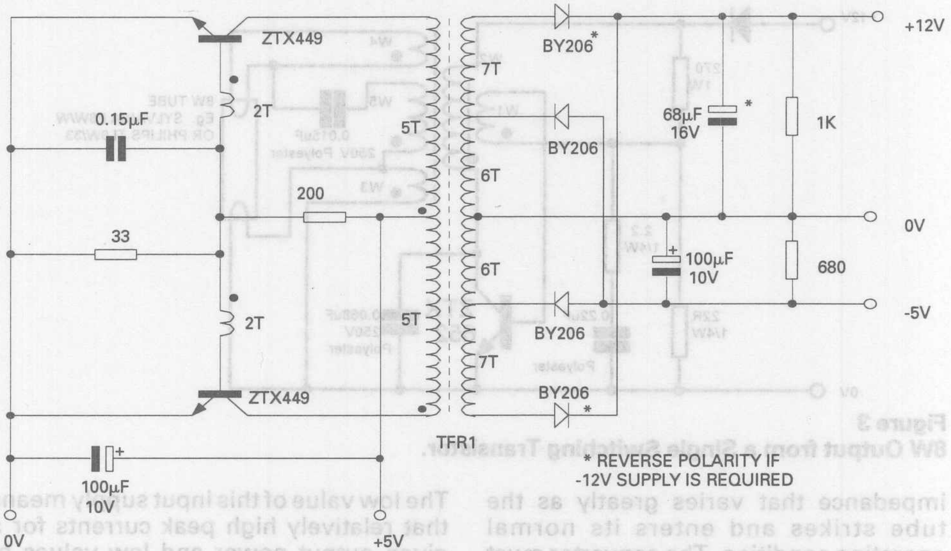


Figure 4
ZTX652 Switching waveforms for 8W Inverter.
Upper Trace - V_{CE} 10V/div
Lower Trace - I_E 1A/div



* REVERSE POLARITY IF
-12V SUPPLY IS REQUIRED

Figure 5
Simple 5W Inverter to Provide +12V and -5V.

25kHz, two Zetex ZTX449 E-Line transistors alternately pass currents up to 1.5A. The 250mV $V_{CE(sat)}$ typically given by these devices gives this simple circuit its high efficiency and good regulation. For even higher output powers, ZTX649 transistors could be substituted.

Capacitor charging is a common application for DC-DC converters. This type of converter appears in flash guns, ignition units, etc. Capacitors are awkward components to charge efficiently because the voltage across them varies greatly during the charging cycle (0 to 400V in a typical flash gun). Flyback converters are particularly suited to this task since they isolate the characteristics of the load from the converter switching element.

However, for a given power output, the peak switching current required in a flyback converter is generally greater than that for a forward converter. So the high current ratings and low saturation voltages of the Zetex E-Line products can be even more important. Figure 6 shows a simple Xenon tube flash circuit based on a flyback converter. The circuit includes a special shut-down sensor which both ensures that the output capacitor is not overcharged and also increases battery life.

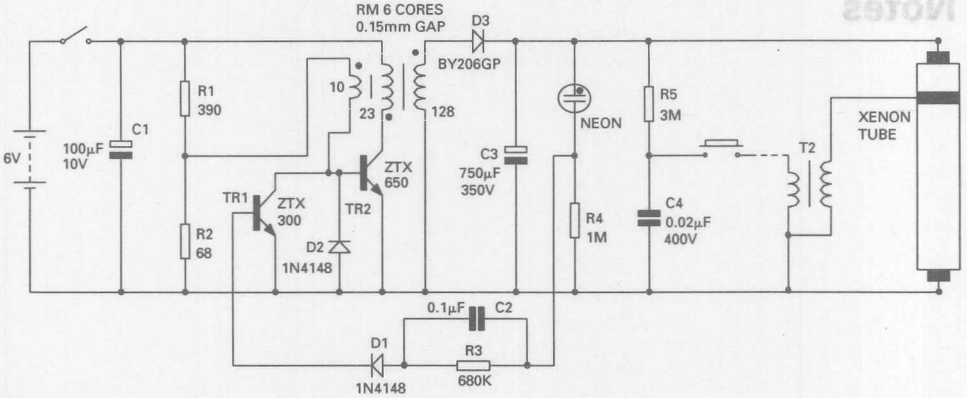


Figure 6
Xenon Tube Flash Circuit with Shut Down to Increase Battery Life.

Summary

DC-DC converters are to be found in a very wide range of electronic products. The type of circuits used varies greatly to suit each application area. All these circuits require switching devices, and the better the switch, the more efficient and smaller the converter can be. Devices possessing fast switching characteristics, require low drive power (high gain), exhibit low voltage drops and can handle high peak currents are ideal for these applications. Modern chip designs and superior packaging allow the Zetex E-Line range of bipolar and MOSFET transistors to meet these requirements for most applications up to 10W. **Please refer to Note 1 below.**

For more detailed information regarding the Zetex transistors featured in this Application Note, please refer to the following publications:

Semiconductor Data Book

Book 1	Through Hole Components
Book 2	High Performance Surface Mount Components

Note 1:

Since the original issue of this application note, Zetex have introduced many new transistor ranges including the ZTX850 NPN series rated to 5A DC, and the complementary ZTX 950 series. These ranges represent the highest current capability of any E-Line /TO92 transistor. High gain or Super-β parts are also available which allow extremely high conversion efficiencies due to very low on-state loss and drive power requirements.

These ranges are detailed in later application notes.

Notes

The ZTX415 Avalanche Mode Transistor

An Introduction to Characteristics, Performance and Applications

Neil Chadderton

Introduction

Avalanche mode devices can provide extremely high switching speeds and are capable of producing current outputs far in excess of that obtained from conventional circuits.

These attributes lend themselves to many applications, including; laser diode drivers for ranging, measurement, and collision avoidance systems, Pockel cell drivers for laser Q switches, SAW device exciters, streak cameras and fast high voltage/ high current pulse generators.

The Zetex Semiconductors ZTX415 is an avalanche transistor that ideally suits this mode of operation and application. The device is available in the proven silicone E-Line package that assures excellent thermal properties to complement the high performance avalanche characteristics. It is also available in the SOT-23 surface mount package (as FMMT415) to enable very low inductance designs.

This Application Note outlines the principle of avalanche mode operation, gives details on important parameter and operating conditions, and suggests a few application circuit examples. (Please refer to Appendix A for a list of reference material).

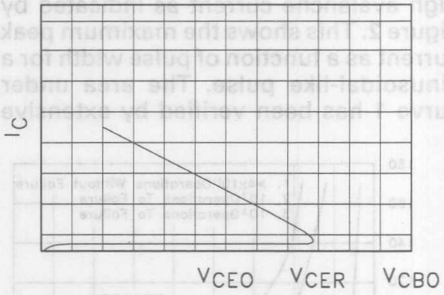


Figure 1
Transistor Characteristics in the Avalanche Region.

What Is An Avalanche Transistor?

Avalanche transistors are characterised by a negative resistance region in their V-I breakdown curve (usually called secondary breakdown) as illustrated in Figure 1. This region permits controlled switching of very high currents in nanoseconds when appropriate external circuitry is employed. The output pulse generated is limited by the primary breakdown B_{VCBO}, the transistor's 'On-state' voltage and the mean dissipation that can be tolerated.

Characteristics

This section reviews the important parameters of an avalanche mode transistor with particular reference to the ZTX415.

The current passed in the secondary breakdown condition, defined as I_{USB}, is dependent on the supply to the device and it's 'On-state' characteristics. The ZTX415 is capable of producing very high avalanche current as indicated by Figure 2. This shows the maximum peak current as a function of pulse width for a sinusoidal-like pulse. The area under curve 1 has been verified by extensive reliability work. For example, devices on a life test have produced current pulses of 60A peak and 10ns pulse width for over 4 x 10^{11} times without failure.

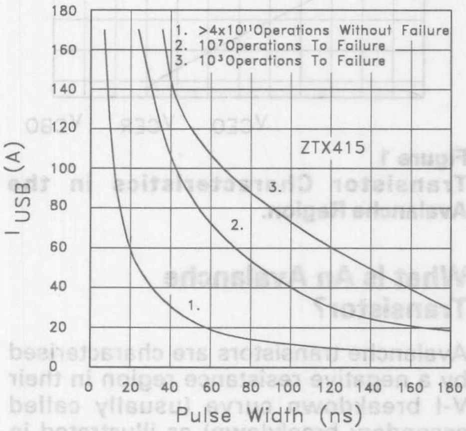


Figure 2
Maximum Avalanche Current v Pulse Width.

The diagram also shows a recorded 10^7 operations to failure (curve 2), and a 10^3 operations to failure, (curve 3, observed sudden failure points). The test circuit used to derive the measurements for these curves is shown in Figure 3, and is basically a single capacitor arrangement.

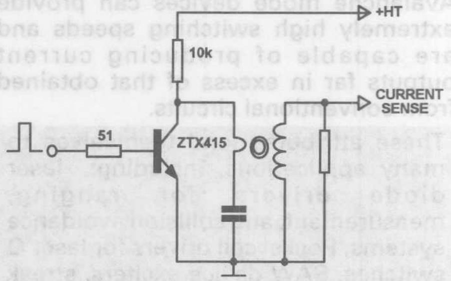


Figure 3
Single Capacitor Test Circuit for I_{usb} Measurements.

The avalanche current is again illustrated in Figure 4 for a typical device, showing its temperature dependency for the specification conditions.

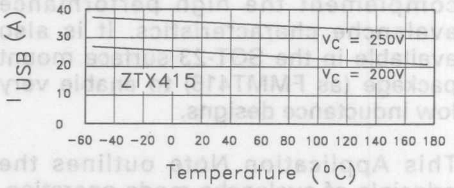


Figure 4
I_{USB} v Temperature.

When using avalanche devices, attention to the base bias and therefore the possible gain variation can be important. Figure 5 shows the gain variation versus collector current over the operating temperature range for the ZTX415.

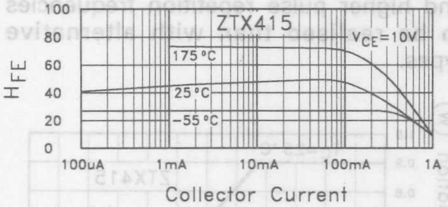


Figure 5
h_{FE} v Collector Current.

The circuit techniques employed, and ultimately the usable operating range, depend somewhat on the static voltage characteristic of the avalanche device, e.g. to obtain a high voltage pulse the transistor must have a high B_{VCES}. The ZTX415 has been designed using high voltage technology to produce a device possessing a minimum B_{VCES} of 260V over the specified temperature range (the 25°C minimum actually being 280V). This makes possible the generation of high current pulses with the minimum of circuitry.

Furthermore, the ZTX415 process is very closely controlled to yield a product with a relatively narrow band of B_{VCES} (Viz: 1.2:1 compared with 2:1 for conventional high voltage product) - an important attribute when designing high voltage stacks employing many devices in series, to produce kilovolt range output pulses. This feature also releases the customer from expensive selections and matching operations, otherwise required to guarantee performance and reliability.

Another important parameter is the minimum voltage required for avalanche operation, below which the device has the switching characteristic of a device in non-avalanche mode. This 'starting' voltage is dependent on the external circuitry employed with the device, but for a simple single capacitor arrangement is seen to vary as shown in Figure 6.

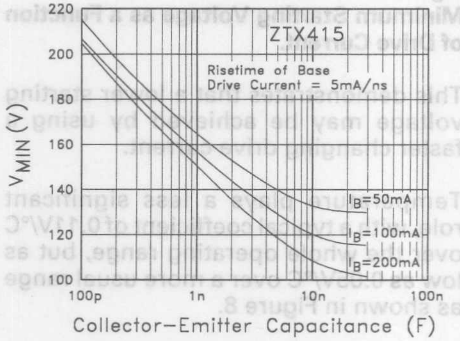


Figure 6
Minimum Starting Voltage as a Function of Capacitance.

This shows the minimum voltage required for avalanche operation as a function of capacitance, for different drive currents. The rate of change of drive current is also relevant, as Figure 7 illustrates.

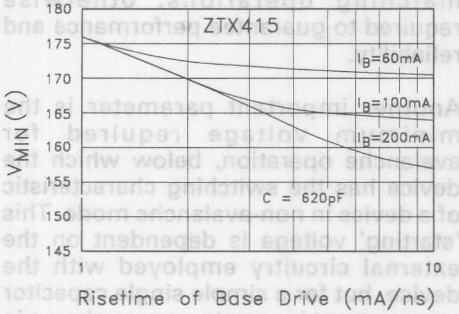


Figure 7
Minimum Starting Voltage as a Function of Drive Current.

This demonstrates that a lower starting voltage may be achieved by using a faster changing drive current.

Temperature plays a less significant role, with a typical coefficient of 0.11V/°C over the whole operating range, but as low as 0.05V/°C over a more usual range as shown in Figure 8.

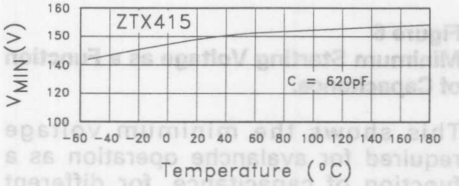


Figure 8
Minimum Starting Voltage as a Function of Temperature.

Any device is eventually limited by the maximum permitted power dissipation allowed for a particular package. However, due to the superior thermal performance of silicone packaging, Zetex devices out-perform any other TO92 style packages available. Referring to Figure 9 will show that the ZTX415 is usable without heatsinking up to an ambient temperature of 175°C. This enables higher currents (or voltages), and higher pulse repetition frequencies to be realised than with alternative types.

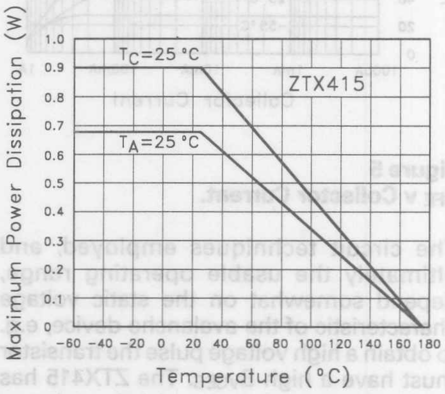


Figure 9
Dissipation De-rating Curves.

Reliability

Extensive reliability research has been performed on all of the DC and Avalanche characteristics of the ZTX415, confirming both the power rating and the inherent quality produced by proven device design techniques. Environmental, life test, and high temperature reverse bias (HTRB) trials are carried out routinely to guarantee continuing quality.

Applications

Avalanche mode characteristics can be realised in a wide variety of circuits from simple single capacitor arrangements through shaped pulse circuits, fast monostables and high current/high voltage pulse generators. The following gives some typical application examples.

The delay line pulse generator, shown in alternative configurations in Figures 10a and 10b, is so called because it depends on a charged delay line - the coaxial cable. The circuit produces a rectangular pulse, dependent on the length and characteristics of the cable.

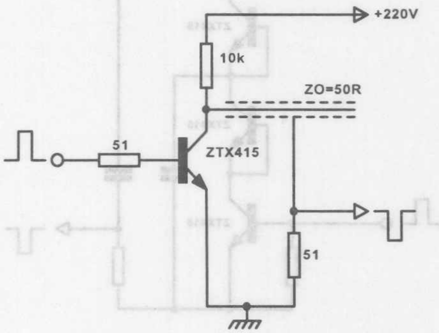


Figure 10a
Delay Line Pulse Generator.

Similar to the delay line generator but using a single capacitor arrangement, is the circuit shown in Figure 11(a). Using a ZTX415, this circuit is capable of producing sinusoïdal-like pulses of tens of amperes, as reproduced in Figure 11(b). This form of circuit is ideal for driving laser LEDs.

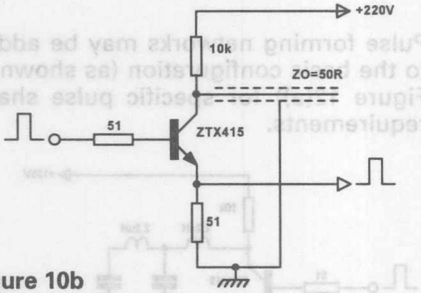


Figure 10b
Delay Line Pulse Generator.

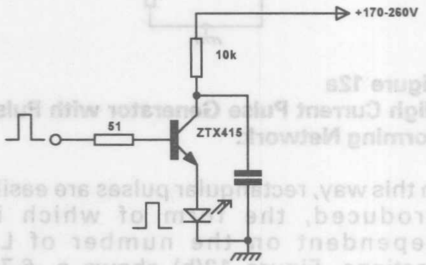


Figure 11a
High Current Pulse Generator for Laser LEDs.

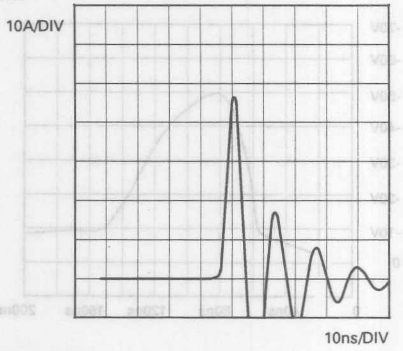


Figure 11b
Current Pulse Generated by Circuit shown In Figure 11a, showing 46A Peak 8ns Wide Sinusoid.

Pulse forming networks may be added to the basic configuration (as shown in Figure 12(a)) for specific pulse shape requirements.

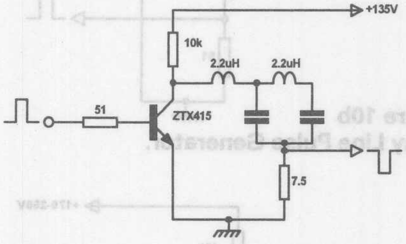


Figure 12a
High Current Pulse Generator with Pulse Forming Network.

In this way, rectangular pulses are easily produced, the form of which is dependent on the number of LC sections. Figure 12(b) shows a -6.7A peak, 90ns pulse generated by the two section generator of Figure 12(a).

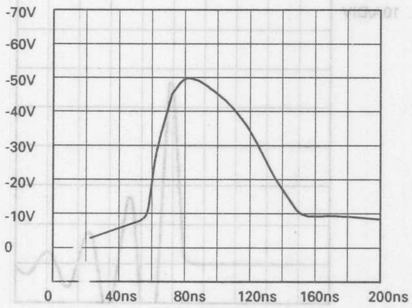


Figure 12b
Current Pulse Generated by Circuit in Figure 12a, showing -6.7A Peak, 80ns Pulse Width.

It is also possible to use the ZTX415 avalanche device in series to allow a higher supply, and therefore generate very high voltage pulses.

Large stacks of avalanche devices are possible as shown in Figure 13, which also shows an additional capacitor included on the lower devices. This is sometimes necessary, depending on application, to ensure avalanche. This particular topology can, by optimising the voltage shared across each transistor, generate voltage steps of many kilovolts. Similar configurations with pulse forming networks are possible as shown in Figure 14.

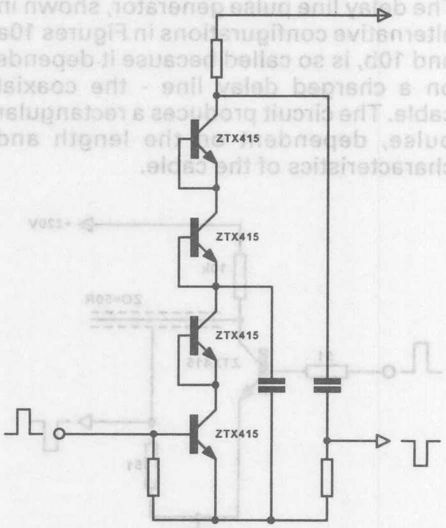


Figure 13
High Voltage Stack (or Series) Operation of ZTX415.

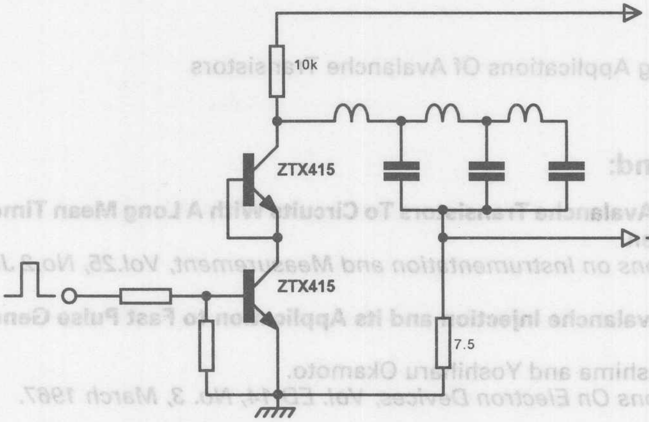


Figure 14
Series operation of ZTX415 with Pulse Forming Network.

When even higher currents are required, it is possible to devise circuitry using parallel ZTX415s as shown in Figure 15. Adjustable biasing makes possible the simultaneous avalanche of the transistor array, allowing current pulses of over one hundred amperes to be produced.

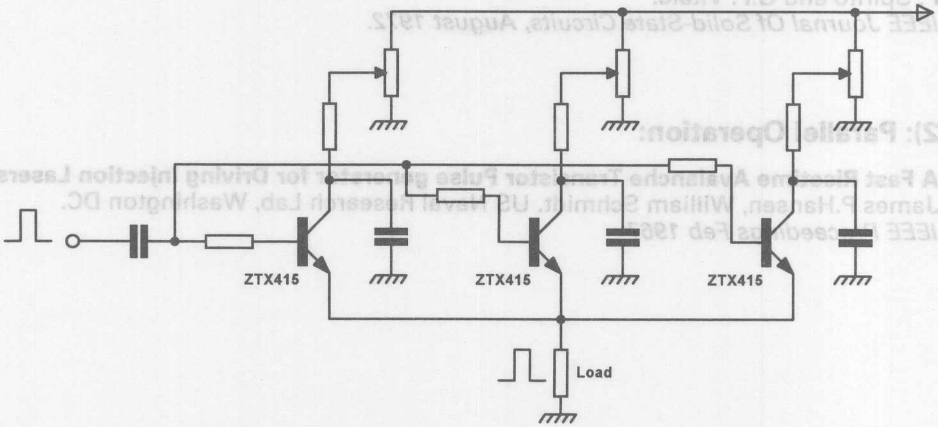


Figure 15
High Current Parallel operation of ZTX415.

Appendix A

Papers Detailing Applications Of Avalanche Transistors

1): Background:

Application Of Avalanche Transistors To Circuits With A Long Mean Time To Failure
Werner B.Herden.
IEEE Transactions on Instrumentation and Measurement, Vol.25, No.2 June 1976.

Properties of Avalanche Injection and its Application to Fast Pulse Generation and Switching
Yoshihiko Mizushima and Yoshiharu Okamoto.
IEEE Transactions On Electron Devices, Vol. ED-14, No. 3, March 1967.

Static and Dynamic Behaviour of Transistors in the Avalanche Region
Paolo Spirito.
Istituto Elettrotecnico, Universita di Napoli, Piazzale Tecchio, 80125 Napoli, Italy
IEEE Journal Of Solid-State Circuits, April 1971.

An Analysis of the Dynamic Behaviour of Switching Circuits Using Avalanche Transistors
P. Spirito and G.F. Vitale.
IEEE Journal Of Solid-State Circuits, August 1972.

2): Parallel Operation:

A Fast Risetime Avalanche Transistor Pulse generator for Driving Injection Lasers
James P.Hansen, William Schmidt. US Naval Research Lab, Washington DC.
IEEE Proceedings Feb 1967

Appendix A (Continued)

3): High Voltage Pulse Generation:

Subnanosecond High-Voltage Pulse Generator
David Brown and Don Martin.
Los Alamos National Laboratory, MS J957, Los Alamos, New Mexico 87545
Rev. Sci. Instrum. 58 (8), August 1987.

Fast High Voltage Pulsers For Nuclear Instruments
Yoneichi Honsono, Satoshi Takeuchi, Ken-ichi Hasegawa, and Masaharu Nakazawa.
Journal Of The Facility Of Engineering, The University Of Tokyo, Vol. XL, No.1 1996.

Avalanche Transistor Pulser For Fast-Gated Operation of Microchannel Plate Image-Intensifiers
Arvid Lundy, James S.Lunsford, and A. Don Martin.
Los Alamos Scientific Laboratory, University Of California, Los Alamos, NM 87545
IEEE Transactions On Nuclear Science, Vol. NS-25, No.1, Feb 1978.

Design of Reliable High Voltage Avalanche Transistor Pulsers
E.Stephen Fulkerson. Lawrence Livermore National Laboratory, Livermore, CA 94551
Rex Booth. Kaiser Engineering Livermore Inc. Livermore, CA 94551

Notes

ZETEX

Application Note 9
Issue 2 January 1996

Zetex Variable Capacitance Diodes

Neil Chadderton

Introduction

The advent of varactor diodes has made a huge impact in many areas of electronic design, which is only too evident in todays consumer products. Formerly, where bulky or unreliable mechanical methods were used, the size, reliability and excellent tracking abilities of the varactor has led to smaller, cheaper and more elaborate circuitry, previously impossible to attain.

Zetex manufacture an extensive range of Variable Capacitance diodes that are processed using ion-implantation techniques to assure accurate doping levels, and hence produce the exacting junction profiles necessary for high performance devices. An overall capacitance range of approximately 1pF to 200pF assures a broad applications base, enabling designs operating from kHz and extending into the microwave region. Product variability within specification is comparable to, or better than, competitors devices; the Hyperabrupt series, for example, are available to 5% tolerance on nominal capacitance, due to close targeting during fabrication. Furthermore, Zetex are capable of matching (either to device type or customer specifications) or banding on capacitance parameters, as required. They are available in surface mount (SOT-23) or E-Line style packages that exhibit low inductance ensuring a wide frequency application, and assure environmental endurance and mechanical reliability.

This application note gives some basic background information, examines the important parameters and specifications for the Zetex range of devices, and suggests a few application circuit examples.

Background

The varactor diode is a device that is processed so to capitalise on the properties of the depletion layer of a P-N diode. Under reverse bias, the carriers in each region (holes in the P type and electrons in the N type) move away from the junction, leaving an area that is depleted of carriers. Thus a region that is essentially an insulator has been created, and can be compared to the classic parallel plate capacitor model. The effective width of this depletion region increases with reverse bias, and so the capacitance decreases. Thus the depletion layer effectively creates a voltage dependent junction capacitance, that can be varied between the forward conduction region and the reverse breakdown voltage (typically +0.7V to -35V respectively for the ZC830 and ZC740 series diodes).

Different junction profiles can be produced that exhibit different Capacitance-Voltage(C-V) characteristics. The Abrupt junction type for example, shows a small range of capacitance due to its diffusion profile, and as a consequence of this is capable of high Q and low distortion, while the Hyperabrupt variety allows a larger change in capacitance for the same range of reverse bias. So called Hyper-hyperabrupt, or octave tuning variable capacitance diodes show a large change in capacitance for a relatively small change in bias voltage.This is particularly useful in battery powered systems where the available bias voltage is limited.

The varactor can be modelled as a variable capacitance (Cjv), in series with a resistance (Rs). (Please refer to Figure 1).

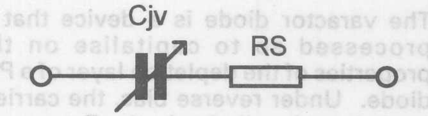


Figure 1
Common Model for the Varactor Diode.

The capacitance, Cjv, is dependent upon the reverse bias voltage, the junction area, and the doping densities of the semiconductor material, and can be described by:

$$Cjv = \frac{Cj\,0}{(1+{}^{Vr}/\varphi\,)^{N}}$$

Where:

Cj0 = Junction capacitance at 0V
Cjv = Junction capacitance at applied bias voltage Vr
Vr = Applied bias voltage
φ = Contact Potential
N = Power law of the junction or slope factor.

The series resistance exists as a consequence of the remaining undepleted semiconductor resistance, a contribution due to the die substrate, and a small lead and package component, and is foremost in determining the performance of the device under RF conditions.

This follows, as the quality factor, Q, is given by:

$$Q = \frac{1}{2\pi f C_{jv}\, R_s}$$

Where:

Cjv = Junction capacitance at applied bias voltage Vr
Rs = Series Resistance
f = Frequency

So, to maximise Q, R_s must be minimised. This is achieved by the use of an epitaxial structure so minimising the amount of high resistivity material in series with the junction.

NOTE: Zetex has produced a set of SPICE models to enable designers to simulate their circuits in SPICE, PSPICE and similar simulation packages. The models use a version of the above capacitance equation and so the model parameters may also be of interest for other software packages. Information is also provided to allow inclusion of parasitic elements to the model. These models are available on request, from any Zetex sales office.

Important Parameters

This section reviews the important characteristics of varactor diodes with particular reference to the Zetex range of variable capacitance diodes.

The characteristic of prime concern to the designer is the Capacitance-Voltage relationship, illustrated by a C-V curve, and expressed at a particular voltage by Cx, where x is the bias voltage. The C-V curve summarises the range of useful capacitance, and also shows the shape of the relationship, which may be relevant when a specific response is required. Figures 2a, 2b and 2c show families of C-V curves for the ZC740-54 (Abrupt), ZC830-6 (Hyperabrupt), and ZC930 (Hyper- hyperabrupt) ranges respectively. Obviously, the choice of device type depends upon the application, but aspects to consider include: the range of frequencies the circuit must operate

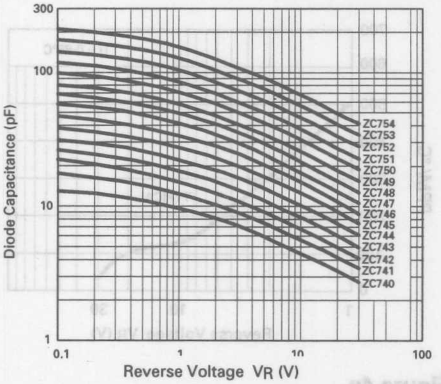


Figure 2a
Typical Capacitance-Voltage Characteristics for the ZC740-54 Range.

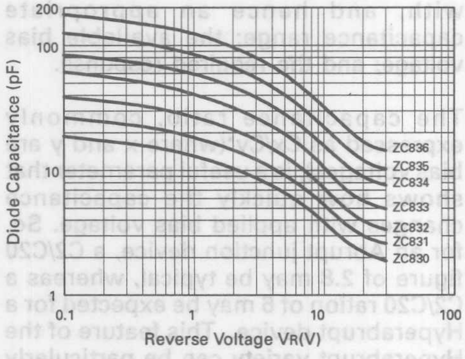


Figure 2b
Typical Capacitance-Voltage Characteristics for the ZC830-6 Range.

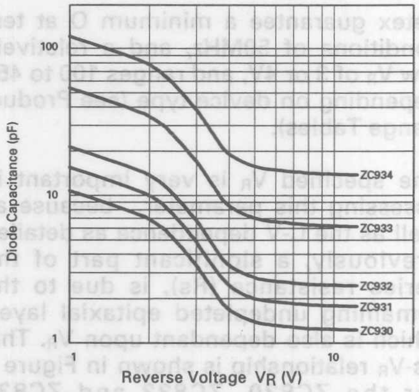


Figure 2c
Typical Capacitance-Voltage Characteristics for the ZC930-4 Range.

with, and hence an appropriate capacitance range; the available bias voltage; and the required response.

The capacitance ratio, commonly expressed as Cx/Cy'(where x and y are bias voltages), is a useful parameter that shows how quickly the capacitance changes with applied bias voltage. So, for an Abrupt junction device, a C2/C20 figure of 2.8 may be typical, whereas a C2/C20 ration of 6 may be expected for a Hyperabrupt device. This feature of the Hyperabrupt variety can be particularly important when assessing devices for battery-powered applications, where the bias voltage range may be limited. In this instance, the ZC930 series that feature a better than 2:1 tuning range for a 0 to 6V bias may be of particular interest.

The quality factor, Q, at a particular condition is a useful parameter in assessing the performance of a device with respect to tuned circuits, and the resulting loaded Q.

Zetex guarantee a minimum Q at test conditions of 50MHz, and a relatively low V_R of 3 or 4V, and ranges 100 to 450 depending on device type (see Product Range Tables).

The specified V_R is very important in assessing this parameter, because as well as the C-V dependence as detailed previously, a significant part of the series resistance (Rs), is due to the remaining undepleted epitaxial layer, which is also dependant upon V_R. This Rs-V_R relationship is shown in Figure 3 for the ZC830, ZC833 and ZC836 Hyperabrupt devices, measured at frequencies of 470MHz, 300MHz and 150MHz respectively, and also serves to illustrate the excellent performance of Zetex Variable Capacitance Diodes at VHF and UHF.

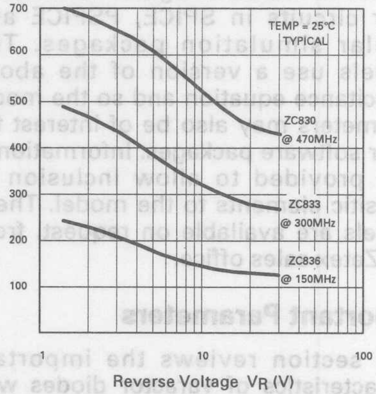


Figure 3
Typical R_S v V_R Relationship for ZC830 Series Diodes.

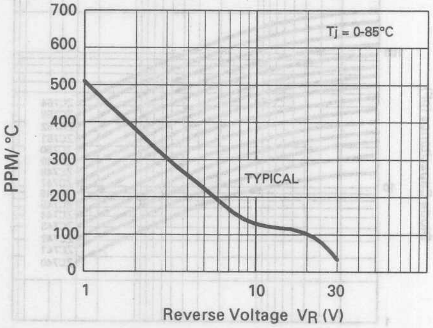


Figure 4a
Temperature Coefficient of Capacitance v V_R for the ZC740 Series.

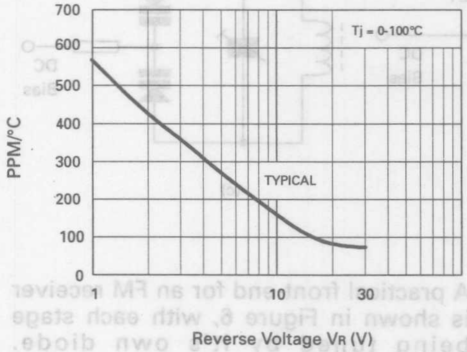


Figure 4b
Temperature Coefficient of Capacitance v V_R for the ZC830 Series.

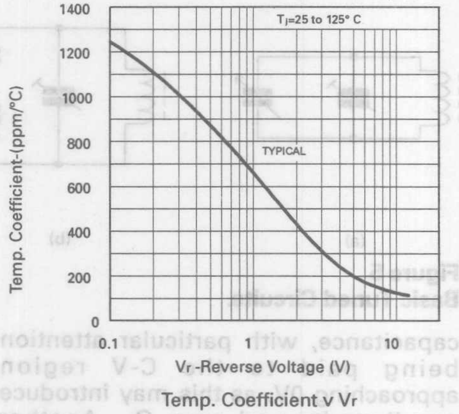


Figure 4c
Temperature Coefficient of Capacitance v V_R for the ZC930 Series.

Also of interest, with respect to stability, is the temperature coefficient of capacitance, as capacitance changes with V_R, and is shown for the three ranges in figures 4a, 4b and 4c respectively.

The reverse breakdown voltage, V(BR) also has a bearing on device selection, as this parameter limits the maximum V_R that may be used when biasing for minimum capacitance. Zetex variable capacitance diodes typically possess a V(BR) of 35V.

The maximum frequency of operation will depend on the required capacitance and the series resistance (and hence useful Q), that is possessed by a particular device type, but also of consequence are the parasitic components exhibited by the device package. These depend on the size, material, and construction of the package. For example, the Zetex SOT-23 package has a typical stray capacitance of 0.08pF, and a total lead inductance of 2.8nH, while the E-Line package shows less than 0.2pF and 5nH respectively. These low values allow a wide frequency application, for example, the ZC830 and ZC930 series, configured as series pairs are ideal for low cost microwave designs extending to 2.5GHz and above.

Applications

Variable capacitance diodes can be used in any tuned circuit application where previously mechanical methods were utilised, and provide a size, cost and performance advantage. This section briefly examines some typical examples of varactor application.

The conventional simple tuned circuit of Figure 5a can also be effected by the varactor version as in Figure 5b, where capacitor C1 isolates the DC bias. The choice of varactor for such a circuit depends on the tuning range and hence

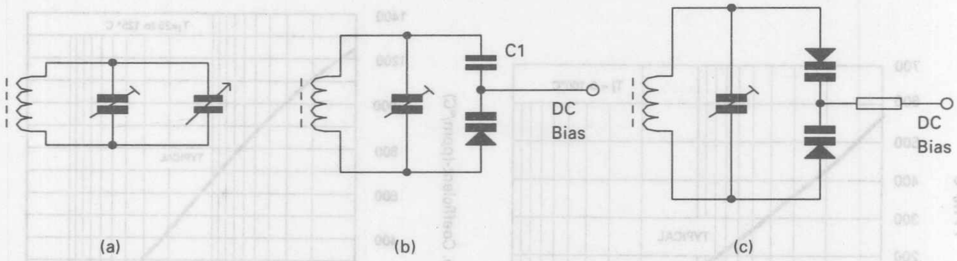


Figure 5
Basic Tuned Circuits.

capacitance, with particular attention being paid to the C-V region approaching 0V, as this may introduce non-linearity and poor Q. Another similar approach is to use the series configuration shown in Figure 5c which, while allowing a lower apparent diode capacitance, also prevents RF rectification at low values of diode bias therefore inhibiting generation of intermodulation products, and also simplifies bias requirements.

A practical front-end for an FM receiver is shown in Figure 6, with each stage being tuned by it's own diode. Multi-stage units are therefore possible without the severe tracking errors, and the massive size penalty inherent to mechanical mechanisms.

As the tuning is now controlled by a voltage, the inevitable inclusion of the microprocessor and memory in many modern receivers has allowed

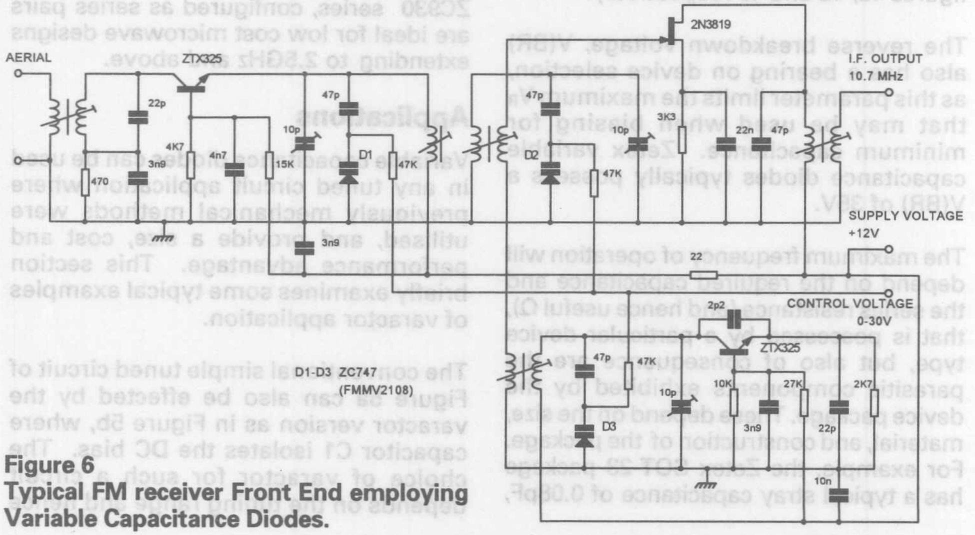


Figure 6
Typical FM receiver Front End employing Variable Capacitance Diodes.

band-scanning and station storage by producing the control voltage automatically. It is noteworthy that the control voltage of any system must possess good voltage and temperature stability.

Perhaps the largest effect on consumer products due to the varactor has been the development of varactor based VHF and UHF tuners for televisions. These have enabled solid state non-mechanical designs that are smaller, more reliable and allow elaborate features such as remote control and station searching. Figure 7 shows a typical circuit of a UHF tuner incorporating varactor diodes. Stage matching is effected by the bias trimmers RV1-5, and allows adjustment remote from the actual tuning element; the mechanical equivalent being to add padding capacitance, or to bend the vanes on an air-spaced capacitor.

Such a tuner can, using the large capacitance range of Hyperabrupt varactor diodes, tune the whole channel range of bands IV and V (470MHz-850MHz).

Another common application for the varactor is as the frequency controlling element in a Voltage Controlled Oscillator (VCO). There are many applications for such circuits, either as stand alone units or as part of a phase locked loop in a frequency synthesiser. This latter method is commonly utilised in radio telecommunications and for the tuning stages in satellite receivers. Closely allied to these are functions such as frequency pulling on crystal oscillators, narrow band FM and temperature compensation of frequency within an oscillator, all of which can benefit from a varactor diode based design.

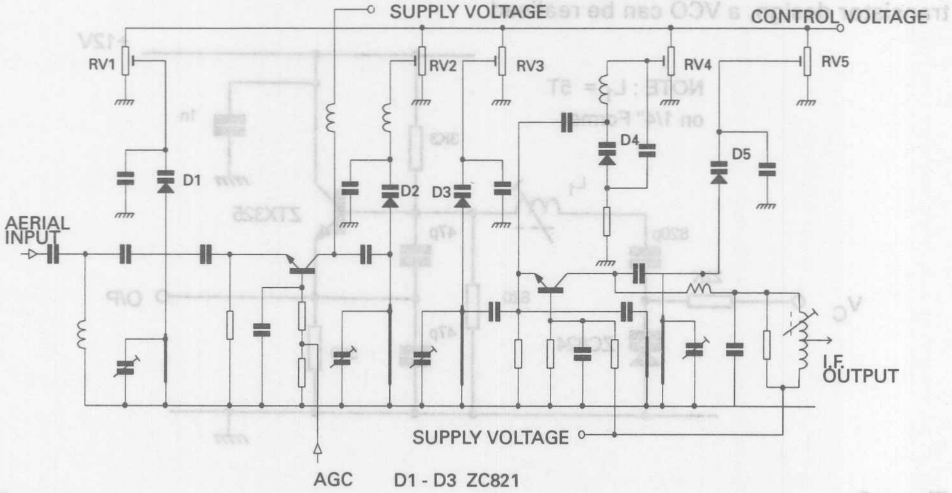


Figure 7
Typical UHF Variable Capacitance Diode Tuner Module.

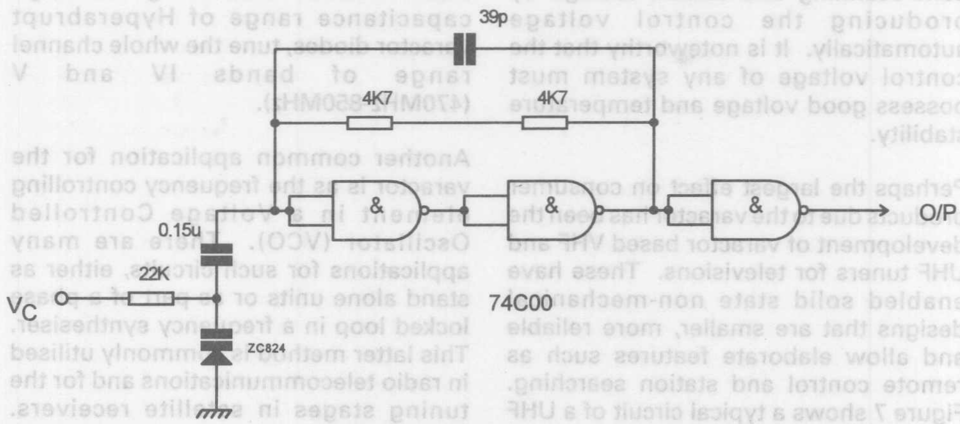


Figure 8
Basic Form of VCO using Logic Gates.

In it's basic form, and for a square wave output, a common method of providing a VCO is to use logic gates coupled as shown in Figure 8; this particular design giving a 1 to 1.25MHz range. For a transistor design, a VCO can be realised by modifying the Clapp configuration as shown in Figure 9. For this example, the frequency can be varied over a 75 to 150MHz range for a 0-30V control voltage.

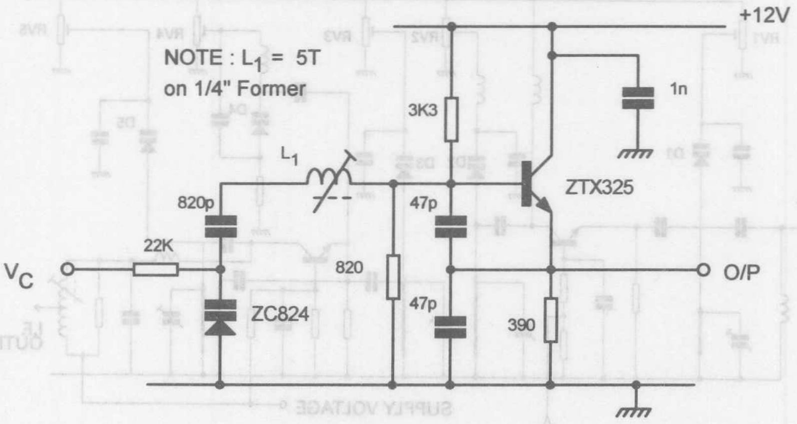


Figure 9
Transistor Effected VCO using Clapp Configuration.

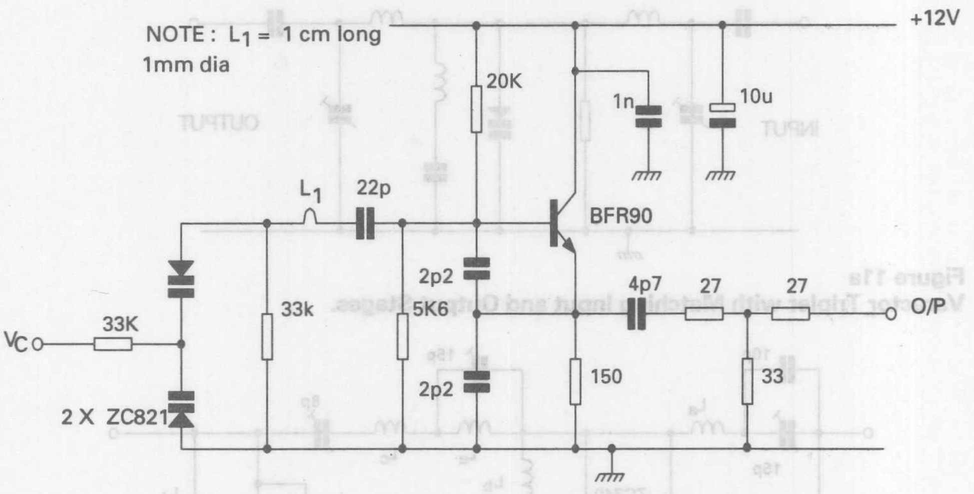


Figure 10
1GHz VCO Producing -5dBm with a 10dB PAD into 50Ω.

Figure 10 shows a similar configuration for a 1GHz VCO. Obviously at this frequency, circuit construction is critical and capable of producing large tuning range changes. For this example, the transistor was mounted in a slot in a small ground plane configured board, and the other components supported by short leads. This produced a signal level of -5dBm with a 10dB PAD into 50 ohms. The second harmonic was observed at -35dB down from the fundamental.

Other techniques using lumped components and Zetex variable capacitance diodes are capable of output at 1.5-2.5GHz, typically for tuning units for satellite receivers; one such design is included in Plessey's *"Satellite, Cable and TV Handbook"* as an alternative to the costly commercial VCO modules.

Low Phase Noise Capability

Due to the geometric features employed in the Zetex range of variable capacitance diodes, these products exhibit extremely low values of leakage current (typically less than 200 pA, and less for low capacitance options) which enables excellent low phase noise performance within VCXO circuitry.

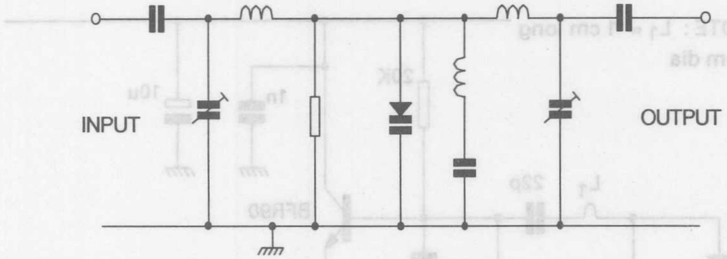
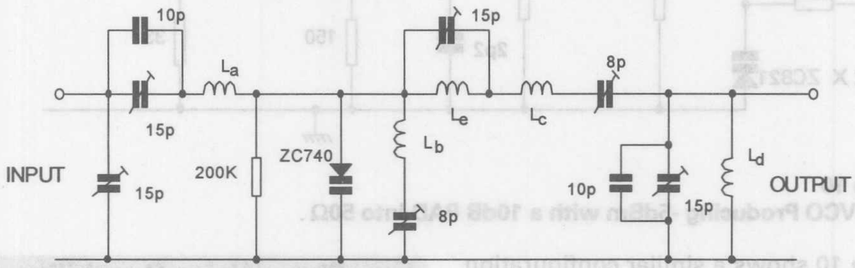


Figure 11a
Varactor Tripler with Matching Input and Output Stages.



Figures11b
Varactor Tripler with Bandpass Filtered Output and a Trap for the Fundamental.

Coil winding data:	
La 5¼ turns 16swg tinned copper ½" dia.	Lc 3½ turns 16 swg tinned copper ⅜" dia.
Lb 3½ turns 18 swg tinned copper ⅜" dia.	Ld ½ turn 16 swg tinned copper ½" dia.
	Le 7 turns 18swg tinned copper ⅜ "dia.

The varactor diode also enables frequency multipliers to be produced that exhibit very high conversion efficiency, a zero DC power requirement, and low component count. Figure 11a shows the general appearance of a varactor tripler, and consists of input and output matching, and a trap for the unwanted second harmonic. Figure 11b shows a similar circuit for a 100-300MHz tripler using a ZC740, and includes a bandpass filtered output and a trap for the fundamental.

Appendix

Zetex Variable Capacitance Diode Product Range.

The tables presented within this Appendix illustrate the standard abrupt, hyperabrupt and hyper-hyperabrupt ranges available with respect to datasheet characteristics and package style. In addition, Zetex can also supply to competitors' type numbers and customers' specific requirements.

HIGH PERFORMANCE HYPERABRUPT

Package		Nominal Capacitance(pF) (V_R=2V, f=1MHz)	Cap. ratio C2/C20 (f=1MHz)		Figure of Merit, Q Minimum (V_R =3V, f=50MHz)
E-Line	SOT 23		Min.	Max.	
ZC820	ZC830	10	5 (4.5)	6.5 (6)	300
ZC821	ZC831	15	5 (4.5)	6.5 (6)	300
ZC822	ZC832	22	5	6.5	200
ZC823	ZC833	33	5	6.5	200
ZC824	ZC834	47	5	6.5	200
ZC825	ZC835	68	5	6.5	100
ZC826	ZC836	100	5	6.5	100
		See note 1	See note 2		

HIGH PERFORMANCE HYPER-HYPERABRUPT

Package	Nominal Capacitance (pF)			Minimum Cap. ratio C1/C4 (f=1MHz)	Figure of Merit Q Minimum (V_R=4V, f=50 MHz)
SOT23	Min.	Max.	at V_R		
ZC930	4.3	5.5	2.5	3	350
ZC931	6.5	7.8	2.5	3.6	300
ZC932	8.5	10.5	2.5	3.1	200
ZC933	18	27	2.5	3.5	150
ZC934	40	65	2.5	3.8	80

Note 1: Available with 5% (suffix B), 10% (suffix A) and 20% tolerance on nominal capacitance.
2: SOT-23 figure in parenthesis when different.

HYPERABRUPT

Package SOT-23	Nominal Capacitance (pF) (f=1MHz)			Cap.ratio C3/C25 (f=1MHz)		Figure of Merit, Q Minimum (V_R=3V, f=50MHz)
	Min.	Max.	at V_R	Min.	Max.	
BBY31	1.8	2.8	25	5 (typ)		-
BBY40	26	32	3	5	6.5	-
	4.3	6	25			-
FMMV105G	1.8	2.8	25	4	6	350
FMMV109	26	32	3	5	6.5	250
FMMV3102	20	25	3	4.5	-	300

ABRUPT

Package		Nominal Capacitance (pF) (V_R=4V, f=1MHz)	Minimum Cap. ratio C2/C30 (f=1MHz)		Figure of Merit, Q Minimum (V_R=4V, f=50MHz)
E-Line	SOT-23		E-Line	SOT-23	
ZC740	FMMV2101	6.8	2.7	2.5	450
ZC741	FMMV2102	8.2	2.7	2.6	450
ZC742	FMMV2103	10	2.7	2.6	400
ZC743	FMMV2104	12	2.8	2.6	400
ZC744	FMMV2105	15	2.8	2.6	400
ZC745	FMMV2106	18	2.8	2.7	350
ZC746	FMMV2107	22	2.8	2.7	350
ZC747	FMMV2108	27	2.8	2.7	300
ZC748	FMMV2109	33	2.8	2.7	200
ZC749		39	2.8		150
ZC750		47	2.8		150
ZC751		56	2.8		150
ZC752		68	2.8		150
ZC753		82	2.8		100
ZC754		100	2.8		100
		See note 3			

Note 3: 10% tolerance on nominal capacitance.

ZETEX

Application Note 10
Issue 2 March 1996

Automotive Relay Drivers using the ZVN4206AV

Protection Free Load Interfacing with Avalanche Rated MOSFETs

David Bradbury

Introduction

The ZVN4206AV is considered to be the ideal automotive relay driver. It allows significant simplification and cost reductions to be made when replacing standard drivers such as bipolar Darlington transistors. The following note explains the popularity of relays as high current switches in automobiles, some of the problems to be solved in designing driver circuits, a standard driver circuit and an improved ZVN4206AV based circuit. The excellent performance of the ZVN4206AV is demonstrated and circuit options discussed.

Automotive Relays

Despite advances in intelligent power semiconductors over the past few years, relays are still extensively used to operate high current automotive loads. Their low conduction resistance, low cost and excellent fault tolerance make relays hard to better in harsh automotive environments. Relay coil currents vary considerably with application but the largest class of automotive relays have 150 to 220mA coils. Controlled by standard logic or similar low power circuits, these relays need a medium power driver transistor to interface to them.

Selection of a suitable driver transistor requires many constraints to be evaluated. Automotive power supplies are rarely clean. Normal operating voltages are from 10 to 15V for normal running but if the relay must remain energised during starter operation then the lower limit can fall to 6V or less. Commercial emergency starting equipment can force the battery voltage beyond 24V so automotive systems must be designed to withstand this voltage for up to 3 minutes to meet present automotive regulations. Load dump and other high energy supply voltage transients must also be withstood. Additional transients are also generated by the relay coil as its magnetic field collapses during turn-off. Wide variation in ambient temperatures must also be taken into consideration. Behind the vehicles instrument panel, ambient operating temperatures can vary from -40°C to over 85°C, whilst within the engine compartment, temperatures can be even higher.

Meeting these constraints usually results in specifying an oversized relay driver or one requiring many protection components.

Standard Drivers

Bipolar Darlingtons are currently the most popular automotive relay drivers as they have the current gain necessary to interface between logic and relay with a single stage circuit. Figure 1 shows a typical Darlington driver with the extra protection components needed to ensure reliability. Diodes provide reverse supply protection and a clamp for the relay's turn-off transient. A power VDR is used to limit positive transients to within the Darlington's breakdown voltage. The saturation voltage of the Darlington (typically over 1V) causes sufficient power dissipation to eliminate the safe use of inexpensive TO92 types (unless they have a 200°C temperature rating - see Note 1 below) so TO126 or TO220 types are often selected. The resulting circuit is expensive, bulky, awkward to assemble and can be subject to mechanical reliability problems.

[Note1:
The Zetex E-Line package allows operation to 200°C, and the product range includes Automotive specific Darlington devices.]

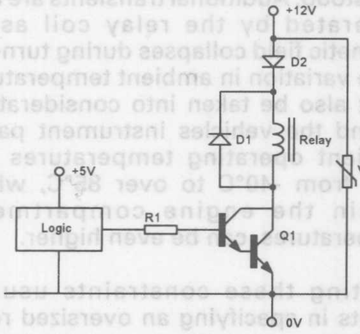


Figure 1
Darlington Driver with Typical Protection Components.

The ZVN4206AV Relay Driver

The introduction of the ZVN4206AV provides a far simpler, if not the simplest possible solution to the problems of relay drivers. It is an avalanche rated 60V 1Ω N-channel MOSFET designed specifically for use with automotive relays. The ZVN4206AV can drive relays with coil currents up to 600mA, and doesn't require a catch-diode to clip the turn-off transient caused as the relay is deactivated. The energy stored in the relay's magnetic field is dissipated harmlessly by avalanche breakdown of the ZVN4206AV's body diode. The same body diode also protects the MOSFET in the event of reverse battery connection. Characterised for 5V gate drive it can be driven directly from standard logic with no interface components.

Figure 2 shows how simple a relay driver circuit can be made by using the ZVN4206AV. The device's MOS gate input requires no DC input current so direct connection to logic is possible without buffers or current setting resistors.

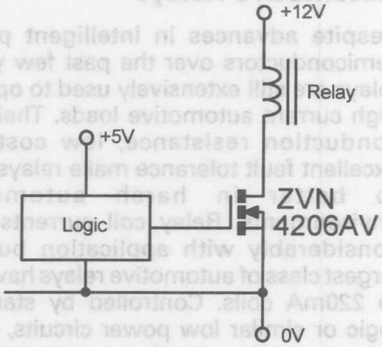


Figure 2
Reduced Component Count Relay Driver using the ZVN4206AV.

When on, the typical relay coil current of 200mA will cause less than 40mW dissipation in the MOSFET at 25°C ambient. Giving an on-voltage drop of only 200mV at this current, the ZVN4206AV will operate the relay with battery voltages 1V lower than can be achieved with Darlington based circuits.

Approximately 2mJ is stored in the magnetic field of the relay whilst energised. When the relay is turned off, the current flowing in the relay's coil causes the drain voltage of the ZVN4206AV to rise rapidly up to the breakdown voltage of the transistor (approximately 80V) and the stored energy is dumped harmlessly into the body diode of the MOSFET. Figure 3 shows the turn-off voltage and current waveforms of a typical relay driver circuit.

If a commercial starting aid is used which doubles the battery voltage, the ZVN4206AV will be forced to pass twice its normal operating current. For loads of 300mA or less it will stand this indefinitely - more than can be said for the relay.

Reverse connected supplies will cause the body diode of the MOSFET to be forward biased and hence conduct. The current will be restricted by the relay's coil resistance to a safe level though the relay will be energised. With bipolar driver transistors, harm can come to the control logic due to a possible current path from a reverse connected battery through the driver to the logic's output. This cannot occur with the ZVN4206AV.

Figure 4 shows a load dump supply transient, caused by the vehicle's alternator if a battery connection should fail during heavy charging. This transient could occur when the relay is on or off and the stress placed on the driver circuit can therefore be very different. If off, the 65V transient will not reach the breakdown voltage of the ZVN4206AV and so no current will flow.

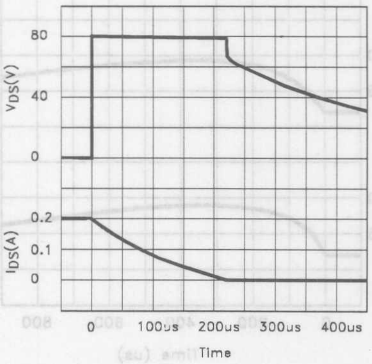


Figure 3
V_{DS} & I_{DS} of ZVN4206AV During Turn-Off of Relay.

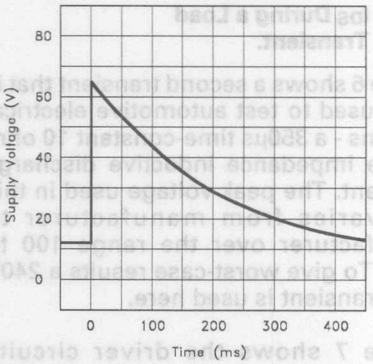


Figure 4
Load Dump Transient Waveform.

Figure 5 shows the current and voltage waveforms that occur should the relay be energised during a load dump transient. By integrating the current-voltage product, it can be calculated that a junction temperature rise of approximately20°C is caused by the transient but this will quickly decay away. Consequently, infrequent load dump transients cause no problems for the ZVN4206AV relay driver circuit.

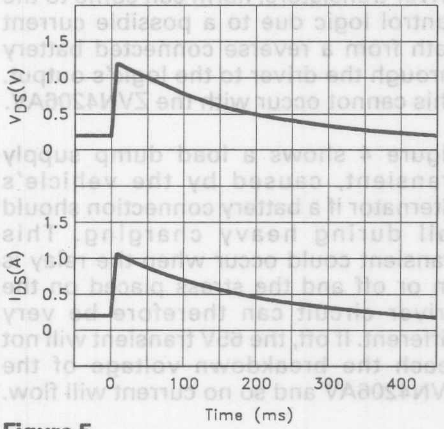


Figure 5
V_{DS} & I_{DS} During a Load Dump Transient.

Figure 6 shows a second transient that is often used to test automotive electrical systems - a 350µs time-constant 10 ohm source impedance inductive discharge transient. The peak voltage used in this test varies from manufacturer to manufacturer over the range 100 to 240V. To give worst-case results a 240V peak transient is used here.

Figure 7 shows the driver circuits response to this transient when the relay is on. The relay coil current rises from its nominal 200mA to a peak of around 900mA and then falls back. The ZVN4206AV maintains a low drain-source voltage for the duration of the transient so the power dissipation caused is negligible.

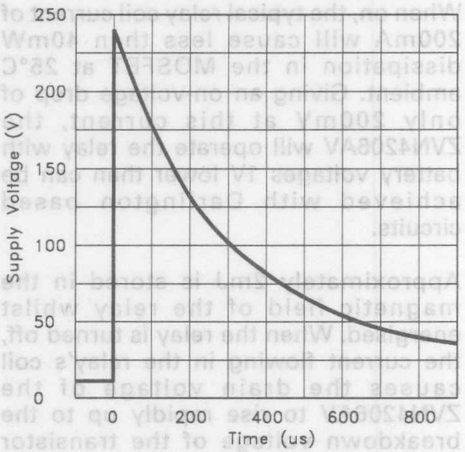


Figure 6
240V Inductive Line Transient Waveform.

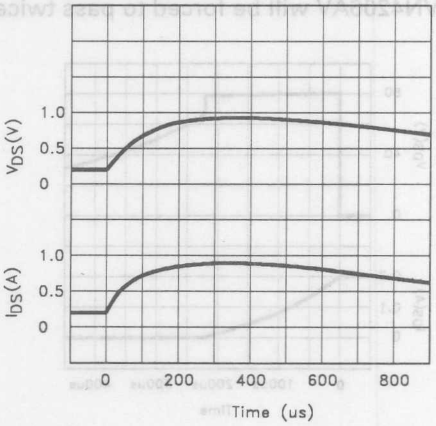


Figure 7
V_{DS} & I_{DS} During a 240V Inductive Line Transient (Device On Before Transient).

Figure 8 shows the circuits response to the 240V transient when the transistor was off before the transient. For this case, the transient forces the ZVN4206AV into avalanche breakdown. The peak avalanche current that flows is limited by the relay's inductance to around 200mA. The energy dumped into the transistor, calculated by integrating the current-voltage product is approximately 5mJ. These stress levels are well within the avalanche capabilities of the ZVN4206AV.

The effect of temperature on the driver circuit should be considered for automotive applications due to their wide operating range. The on-resistance of the ZVN4206AV increases with temperature by about 0.65%/°C and this does reduce its current rating at high temperatures. However it should be remembered that the resistance of the relay's coil also increases with temperature, reducing the driver transistors load current at the rate of 0.43%/°C. Since the power dissipation of the ZVN4206AV is I^2R related, even though the MOSFETs resistance increases with temperature, its overall power dissipation falls as the I^2 factor is more significant. As a result, a ZVN4206AV driving a 200mA relay will operate within its chip temperature ratings for ambients beyond 140°C.

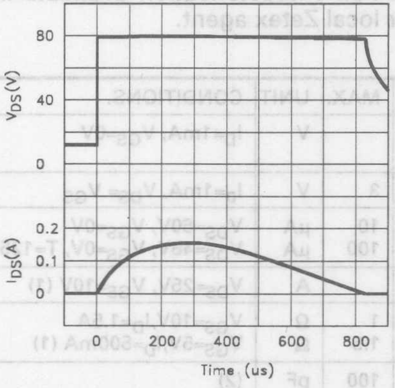


Figure 8
V_{DS} & I_{DS} During a 240V Inductive Line Transient (Device Off Before Transient).

The use of a ZVN4206AV has given a circuit capable of withstanding all commonly applied automotive transients and conditions of misuse whilst eliminating the need for bulky and expensive driver and protection components. This should not only provide worthwhile cost savings over standard drivers but also lead to improvements in reliability due to reduced power dissipation and component count.

Fault Tolerant Relay Drivers

Many automotive relays are mounted in sockets, remote from their drivers. Attempts to clean socket contacts during servicing or fault finding may lead to intermittent shorts being applied to the relay driver and so sometimes this must be catered for. The circuit in Figure 9 shows a ZVN4206AV relay driver circuit which includes load current sensing. In the event of the load current exceeding 300mA, the ZTX108 is turned on, indicating a fault to the controlling microprocessor. The power dissipated in the ZVN4206AV (12V x 1.5A) - 18W can be withstood safely for around 50ms, plenty of time for the microprocessor to recognise the problem and turn off the driver.

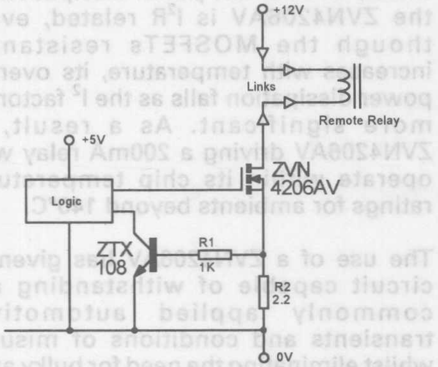


Figure 9
ZVN4206AV Relay Driver Circuit with Overcurrent Protection.

The need for such protection in remote relay systems can be eliminated by putting the driver transistor in the relay module. Figure 10 shows a circuit that can be used to take advantage of this technique. The output resistor in the logic unit protects it against accidental shorts and the inexpensive resistor/zener circuit included in the relay module with the ZVN4206AV gives ESD and open-circuit drive line protection.

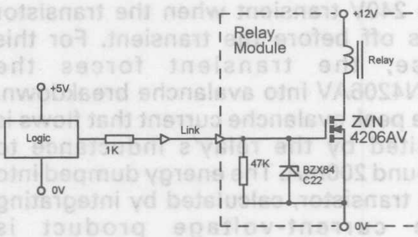


Figure10
ZVN4206AV Relay Driver Module.

Appendix

Partial Characterisation of the ZVN4206AV. Full characterisation available in the Through Hole Data Book available from your local Zetex agent.

PARAMETER	SYMBOL	MIN.	MAX.	UNIT	CONDITIONS.
Drain-Source Breakdown Voltage	BV_{DSS}	60		V	I_D=1mA, V_{GS}=0V
Gate-Source Threshold Voltage	$V_{GS(th)}$	1.3	3	V	I_D=1mA, V_{DS}= V_{GS}
Zero Gate Voltage Drain Current	I_{DSS}		10 100	µA µA	V_{DS}=60V, V_{GS}=0V V_{DS}=48V, V_{GS}=0V, T=125°C(2)
On-State Drain Current	$I_{D(on)}$	3		A	V_{DS}=25V, V_{GS}=10V (1)
Static Drain-Source On-State Resistance	$R_{DS(on)}$		1 1.5	Ω Ω	V_{GS}=10V,I_D=1.5A V_{GS}=5V,I_D=500mA (1)
Input Capacitance	C_{iss}		100	pF	(2)
Avalanche Current - Repetitive	I_{AR}		600	mA	
Avalanche Energy - Repetitive	E_{AR}		15	mJ	

(1) Measured under pulsed conditions. Pulse width=300µs. Duty cycle ≤2% (2) Sample test.

Application Note 11
Issue 2 October 1995

Features and Applications of the FMMT618 and 619

High Current SOT23 replaces SOT89, SOT223 and D-PAK

David Bradbury

Switch 6A loads using a SOT23 transistor?

Zetex has developed this new range to meet ever increasing demands for higher performance in smaller packages.

At first sight the performance of the FMMT618 and FMMT619 transistors may appear unbelievable for SOT23 packaged devices, however the following note describes how this has been achieved. The note also discusses a number of applications where they are ideally suited, which until now were often forced to utilise the larger SOT89 or SOT223 packaged devices. Figure1 provides a graphic example of this performance.

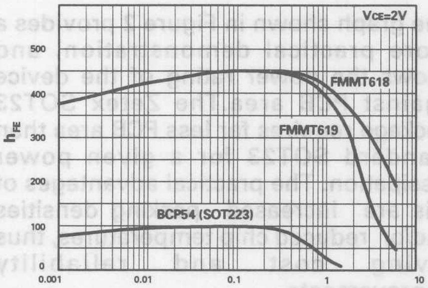


Figure 1
Comparison of h_{FE} profiles.

Main Product Features

- * V_{CEO} up to 50V
- * I_{CM} up to 6A
- * $I_{C\,CONT}$ up to 2.5A
- * P_{tot} of 625mW
- * High h_{FE} - e.g 300 min @ 200mA
- * Very low $V_{CE(sat)}$ - 200mV max @ 2.5A

Comparison with SOT89 and SOT223 Parts

Table 1 gives a quick comparison of the performance of the Zetex FMMT618 & FMMT619 with two industry standard SOT89 and SOT223 parts.

	ZETEX product		Industry Standard	
Package	FMMT618 SOT23	FMMT619 SOT23	BCX54 SOT89	BCP54 SOT223
BV_{CEO}	20V	50V	45V	45V
I_C	2.5A	2A	1A	1A
$I_{C(max)}$	6A	6A	1.5A	1.5A
$h_{FE(min)}$ @ I_C	200 2A	200 1A	25 0.5A	25 0.5A
$V_{CE(sat)}$max @ I_C	50mV 1A	200mV 1A	500mV 0.5A	500mV 0.5A
P_{tot} (mW)	625	625	1000	1500

Table 1
Parametric Comparison.

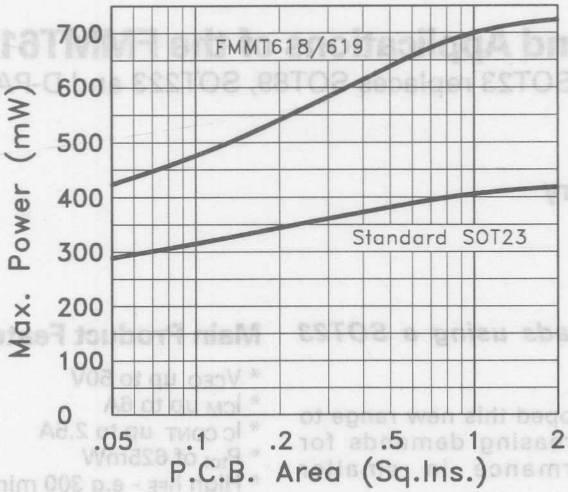


Figure 2
Practical P_D vs PCB area (FR4).

No other manufacturer in the world offers similar performance parts to the FMMT618 and FMMT619 in SOT23. Even in the larger SOT89 and SOT223 packages, few devices are available that can match the current handling and low saturation voltage of the new Zetex range. This level of performance has been achieved by a unique combination of package and chip development.

The Zetex SOT23 Package

The FMMT618 SOT23 series is capable of dissipating 625mW measured on the industry standard 15 x 15mm ceramic substrate.

Most other manufacturers offer power ratings of half this level. The key factor in obtaining such a high power rating is minimising the thermal resistance between the back of the chip and the solder points. For a standard SOT23 package, this thermal resistance is of the order of 280°C/W. Using a custom designed lead frame and multi-metal plating system Zetex have been able to reduce this to only 100°C/W. On an infinite heatsink (say if the test substrate is clamped to a fan cooled heatsink) the package could actually dissipate 1.25W safely but such facilities are rarely available.

The graph shown in Figure 2 provides a more practical demonstration, and shows the power rating of the device against PCB area.The Zetex SOT23 package requires far less PCB area than standard SOT23 for a given power dissipation. The practical advantages of this are increased packing densities and/or reduced chip temperatures, thus giving cost and reliability improvements.

Zetex Matrix Chip Technology

The saturation voltage of the FMMT618 is typically only 130mV at 2.5A and h_{FE} remains useful up to its peak current limit of 6A.

Two major feature enhancements have contributed to the very impressive performance of the FMMT618 and FMMT619 transistors.

Firstly a custom designed SOT23 frame has allowed an extremely large chip to be encapsulated ie. 34 x 36 thou. Secondly and more importantly, the Zetex pioneered 'Matrix' chip design has been used.

In a matrix transistor it is possible to utilise virtually all the die as active area yet minimise the distributed base resistance, ensuring all the area works effectively.

The low saturation voltage obtained is both useful to the circuit designer in maximising circuit efficiency and power applied to the load, whilst also extending current ratings. For most SOT23 packaged bipolar devices, the maximum current capability is defined by power rating and saturation voltage. By optimising both of these characteristics Zetex has produced a range of devices that outperforms many SOT89 and SOT223 parts.

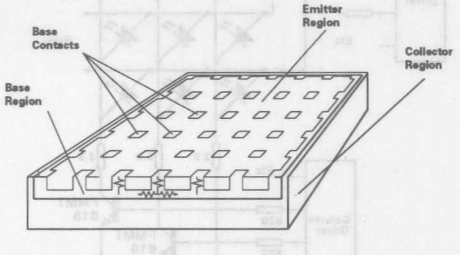


Figure 3
The Matrix Geometry
Distributed base resistance is minimised using a large matrix of base contact holes. By keeping the size of these holes small, little emitter area is lost and so active chip area is maximised.

Replacing these larger package types with the low saturation voltage FMMT618 and FMMT619 transistors will reduce power dissipation, improve circuit efficiency and reliability, while saving on PCB area and component costs.

FMMT618 Applications

Designed with heavy duty battery powered applications in mind, the 20V FMMT618 handles 2.5A continuously, giving a typical saturation voltage of only 130mV at this current. With a minimum h_{FE} at 2A of 200 and at 6A of 100, this transistor is an ideal driver of heavy loads such as motors, lamps, IR LEDs etc. It is also very useful in saturation voltage critical applications such as supply switching and low voltage DC-DC converters.

In Figure 4 an FMMT618 is being used to drive two ZME50 infra-red LEDs in a domestic remote control link. When a key is pressed on the 8 x 4 keypad, the MV500 control IC generates a sequence of pulses which includes synchronisation and key identity data. These 16ms wide pulses are amplified by the FMMT618 and transmitted by the ZME50 diodes. To maximise range, each photodiode must be pulsed at 1-2A so the combined load current for the FMMT618 is 2-4A.

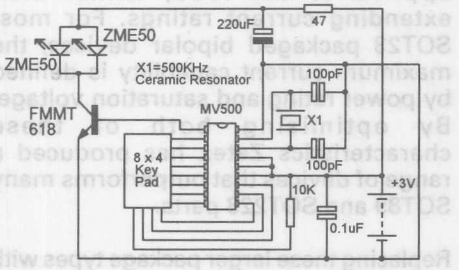


Figure 4
IR Remote Control Transmitter.

Since the remote control unit must be kept small, its power source is just two 1.5V cells giving a 3V nominal supply. To achieve the required LED current, all of this supply must be applied across them so low saturation voltage is essential. Replacing TO126 packaged devices commonly used for this application the FMMT618 provides lower saturation voltages, lower component costs and reduced size. It also allows more of the PCB to be automatically assembled - reducing assembly costs too.

Figure 5 shows the output stage of an LED based 'Moving Messages' display. These systems use thousands of LEDs so it would be impractical to have individual drivers for each LED. To avoid this, the LEDs are wired in the form of a matrix which ideally reduces the number of drivers required to the square root of the number of LEDs. The penalty for this saving is that the drivers must pass much higher currents. The circuit shown is just a segment of a 32 x 32 matrix display. Each is given an average current of 15mA by pulsing it at 0.5A with a duty cycle of 3.125%. The column drivers must be capable of supplying 0.5A but as they are driven from low power logic shift registers they must have high gain. The FMMT618 transistors work well as column drivers, needing only 5mA of base drive to ensure a saturation voltage below 50mV.

For pagers and other 5V circuits that for size reasons must be powered from a 1.5V cell, the FMMT618 can be used as part of an efficient up-converter. The standard boost converter shown in Figure 6 exploits the fast switching and low saturation voltage of the SOT23 transistor to give up to 300mA at efficiencies over 70%.

Real applications are likely to require lower current but significant board area and efficiency gains are to be made by replacing SOT89 and SOT223 transistors in this application.

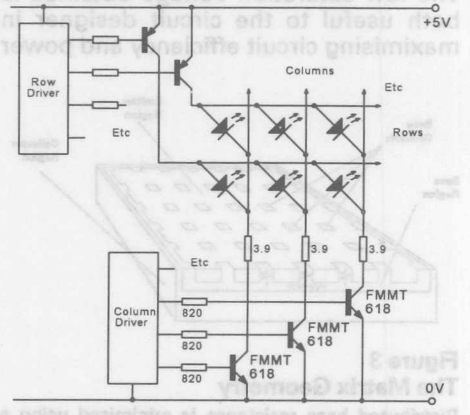


Figure 5
Lamp Matrix.

ZETEX

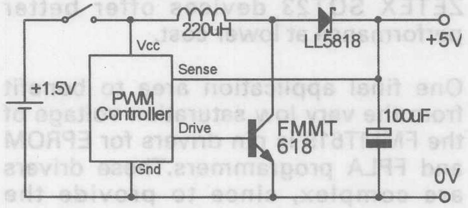


Figure 6
Step-Up converter.

FMMT619 Applications

Intended for higher voltage applications, the 50V FMMT619 still handles a creditable 2A continuous and 6A peak. Giving a typical saturation voltage of just 125mV at 1A for just 10mA of base drive, this transistor can be used to replace inefficient Darlingtons in stepper motor drivers and print head drivers. Also, hybrid DC-DC switching converters where small size and low losses are important can significantly gain by the use of FMMT619 transistors.

Figure 7 shows all that is necessary to use the FMMT619 in a printer's stepper motor driver. Running from a switched 24V-12V supply, the 4-phase motor takes a peak current of around 0.8A. With just 10mA base drive from a CMOS logic buffer the FMMT619 gives a saturation voltage of only 100mV.

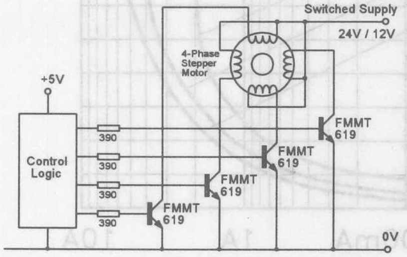


Figure 7
Stepper Motor Driver.

The resulting power loss in the device is so low that the four drivers required can be packed into a PCB area limited by component size rather than power dissipation. Stepping motors can force reverse collector currents through their drivers but the high reverse beta of the FMMT619 allows this tp occur safely, without risk to the transistor or it's driver.

Hybrid DC-DC converter manufacturers are continually working to provide ever higher output power from decreasing package sizes. The push-pull converter shown in Figure 8 is typical of unregulated versions of this type of product and one that can significantly benefit from the use of FMMT619 transistors.

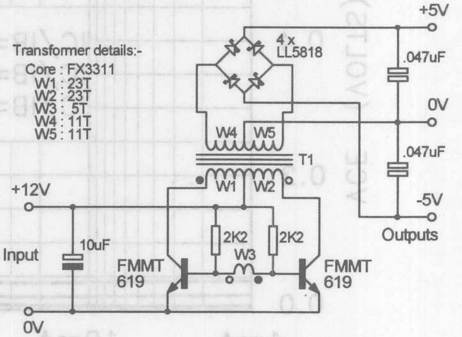


Figure 8
Typical hybrid DC-DC converter.

Taking 12V in and producing ±5V isolated supplies, this self oscillating converter runs at a switching frequency of around 40kHz. The transformer is wound on a small toroidal core and the complete circuit built on a small PCB or ceramic substrate that can be encapsulated in a DIL style package. Conceivably by employing FMMT619 switching transistors this circuit could give a power throughput of over 10W but output rectifier losses and the low power dissipation capability of the small substrates used means a more conservative throughput must be specified. By providing lower saturation voltages and lower thermal resistance the FMMT619 transistors give this standard circuit improved output, load regulation and power.

Higher power converters no longer need use SOT89 or SOT223 types as the ZETEX SOT23 devices offer better performance at lower cost.

One final application area to benefit from the very low saturation voltage of the FMMT619 is pin drivers for EPROM and FPLA programmers.These drivers are complex, since to provide the necessary versatility such programmers must be capable of switching logic signals, supplies and programming pulses to 24-28-40 pin sockets. Despite a small supply current requirement, low saturation voltage is vital. Figure 9 shows the typical saturation voltage of an FMMT619 against collector current demonstrating how the transistor can be used as a supply switch, introducing only 10-20mV of offset.

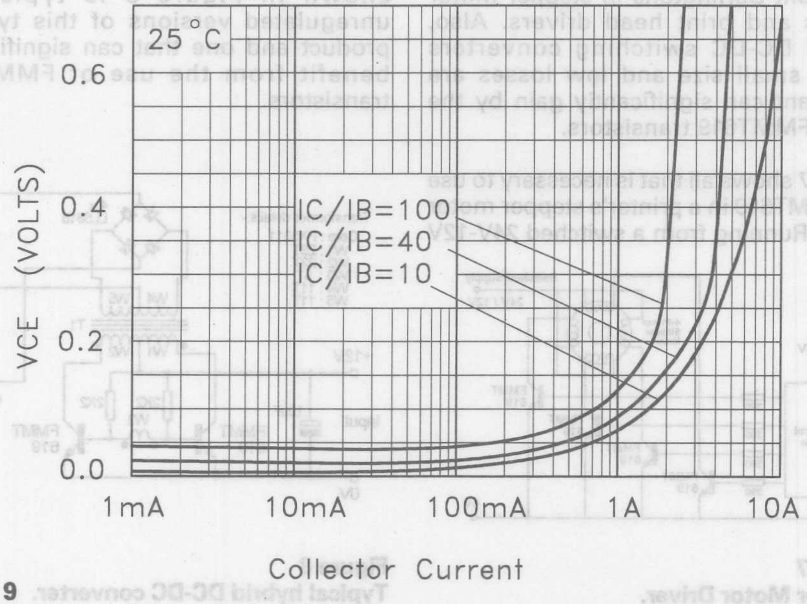


Figure 9
FMMT619 $V_{CE(sat)}$ vs I_C.

ZETEX

Application Note 12
Issue 2 January 1996

The FMMT718 Range, Features and Applications

Replacing SOT89, SOT223 and D-Pak Products with High Current SOT23 Bipolar Transistors.

David Bradbury
Neil Chadderton

Designers of surface mount products wishing to drive loads with currents above a few hundred milli-amps were previously forced to resort to using either large through-hole products or expensive SOT89, SOT223 and D-Pak surface mount transistors.

Now with the introduction of the Zetex FMMT718 and FMMT618 series of PNP and NPN bipolar transistors, loads of up to 6A peak, 2.5A continuous can be driven by SOT23 packaged devices.

These SuperSOT devices provide many advantages including vastly improved circuit efficiency, component and board space savings, and improved reliability. It is true to say that these state of the art devices give performance unmatched by any other SOT23 transistor and many larger SOT89 and SOT223 devices.

This note outlines the features, benefits and applications of the Zetex PNP SuperSOT series, plus some additional NPN applications, and follows on from application note AN11 which covered the low voltage variant NPN types FMMT618 and FMMT619.

Since AN11 detailed how Zetex achieved the exceptional current and saturation voltage performance given by the SuperSOT range, package characteristics and chip design will not be repeated in this note.

Features and Benefits of the FMMT618/718 Series

The key parameters of the FMMT618 and FMMT718 series are listed in Table 1 (overleaf).

The first feature to note is that these SOT23 transistors are capable of dissipating 625mW - around twice the industry standard for SOT23 packaged products. Achieved using a custom lead frame, this means that for a given power dissipation a Zetex SuperSOT chip will run cooler than any competitors' SOT23 devices. This gives improved reliability and the option to reduce PCB area if desired.

Secondly, the saturation voltages are the lowest of any SOT23 of comparable BV_{CEO} in the market place today and lower than many competitors SOT89 and SOT223 types. This translates to lower power dissipation in switching applications, again giving improved reliability and smaller PCB areas.

The SuperSOT Series								
POLARITY	NPN				PNP			
	FMMT618	FMMT619	FMMT624	FMMT625	FMMT718	FMMT720	FMMT722	FMMT723
BV_{CEO}	20V	50V	125V	150V	20V	40V	70V	100V
I_C CONT	2.5A	2A	1A	1A	1.5A	1.5A	1.5A	1A
I_C MAX	6A	6A	3A	3A	6A	4A	3A	2.5A
Midband h_{FE}	450	450	450	450	450	450	450	450
Typical h_{FE} at I_C	360 2A	225 2A	140 1A	45 1A	230 2A	290 1A	275 1A	250 1A
Typical $V_{CE(sat)}$ at I_C	130mV 2.5A	150mV 2A	165mV 1A	180mV A	145mV 1.5A	245mV 1.5A	140mV 1A	210mV 1A
P_{tot}	625mW	625mW	625mW	625mW	625mW	625mW	625mW	625mW

*Measured with device mounted on a 15 x 15 x 0.6 mm ceramic substrate.

Table 1
NPN and PNP SuperSOT Series Parametric Overview.

These features combined with high mid-band h_{FE} and fast switching speeds make the series ideal for switching applications such as DC-DC converters, motor drivers, lamp and solenoid drivers, display drivers, power supply line switching, buffers etc. These transistors will also suit linear applications, e.g. the low series base resistance inherent in the SuperSOT design approach gives excellent low noise performance.

A few example applications are suggested below, that exploit many of the key features of the FMMT718/618 series.

'H'-Bridge Motor Drivers

'H'-bridge motor drivers are used in a wide range of products such as disc drives, coin control mechanisms, automotive applications, servo systems, toys etc. These drivers provide bi-directional outputs from single polarity supplies, usually under the control of a logic IC or microcontroller. They are usually constructed using two NPN and two PNP transistors, all operating in grounded emitter mode (see Figure 1). By turning on one NPN device and the diagonally opposite PNP device, (say NPN1 and PNP1) virtually all the supply voltage can be applied across the motor load. Switching the second pair of transistors instead reverses the supply to the load. 'H'-bridge transistors often require additional collector-emitter diodes to protect the drivers from regenerative currents and transients that can be generated by the load motor.

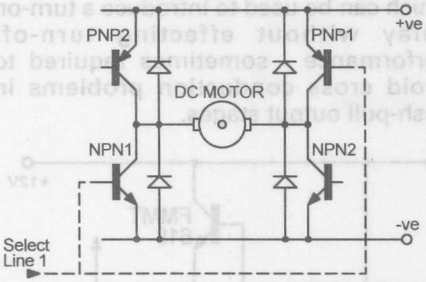


Figure 1
Conceptual 'H'-Bridge Motor Driver.

In battery powered applications it is vital that as much of the supply as possible is applied across the load, maximising battery life through greater efficiency and lower end of life battery voltage. Using the FMMT618 and FMMT718, the bridge circuit shown in Figure 2 will handle load/stall currents up to 1.5A. The circuit can easily be adapted for lower current motors by increasing the value of the base drive resistors. (Set I_B for the PNPs to 1/50 of the maximum load current and I_B for the NPNs to 1/100). The saturation voltage losses at 1.5A are a total of only 0.3V for both NPN & PNP transistors combined, and at lower load currents less than half this level can be expected.

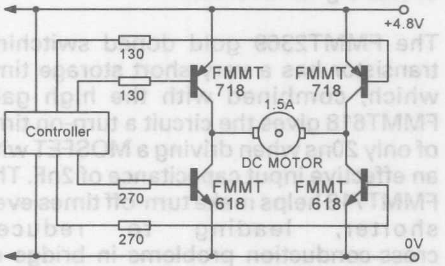


Figure 2
'H'-Bridge Motor Driver using SuperSOT Bipolar Transistors.

The combination of low saturation losses and low base drive requirements of the FMMT618/718 gives improved motor performance and endurance. Parallel diodes are often not necessary for this circuit as the reverse h_{FE} of the driver transistor is sufficiently high to conduct regenerative currents and transients safely away. The small size of the SOT23 package and reduced component count allowed means this bridge circuit can be constructed using much less PCB area than previous designs, giving product size and cost improvements.

For higher voltage motors, the controlling logic is rarely capable of driving the PNP transistors directly. In Figure 3, this problem is overcome by the use of a pair of general purpose NPN transistors to act as buffers/level translators. The h_{FE} of the bridge transistors becomes more important as the supply voltage to the motor is increased as base drive losses are supply dependant. The high h_{FE} of the FMMT618/718 means that these losses can be minimised, saving PCB area and cost.

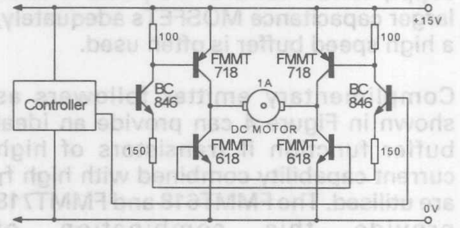


Figure 3
'H'-Bridge Motor Driver with Small Signal Buffer Transistors.

Both of these 'H'-bridge circuits are intended for direct logic drive though the circuit given in Figure 2 requires logic with high current outputs (up to 30mA) if 1.5A loads are to be driven. If this logic drive level is not available, the buffer circuit of Figure 3 provides a solution which is not only inexpensive, but outperforms most SOT89 solutions and many SOT223 based circuits too.

Power MOSFET Gate Drivers

The input capacitance of power MOSFETs and IGBTs range from a few hundred picofarads to tens of nanofarads. When Miller effects are taken into account (the amplification of feedback capacitances) by using the more valid method of evaluating gate charge rather than Ciss to calculate effective input capacitance, values around three times higher are obtained. To minimise switching losses in these power devices, particularly in high frequency converters, it is vital that the gate capacitances are charged and discharged as rapidly as possible. Consequently, driver circuits must act as low impedance voltage sources, capable of supplying large transient charge currents. Since standard switching power supply control ICs are rarely able to drive larger capacitance MOSFETs adequately, a high speed buffer is often used.

Complimentary emitter followers as shown in Figure 4 can provide an ideal buffer function if transistors of high current capability combined with high f_T are utilised. The FMMT618 and FMMT718 provide this combination of characteristics so that the 10nF effective capacitance of two IRF840 MOSFETs in parallel can be charged to 12V in under 30ns - a feat requiring a peak current of around 4A. The circuit includes a resistor which can be used to introduce a turn-on delay without effecting turn-off performance - sometimes required to avoid cross conduction problems in push-pull output stages.

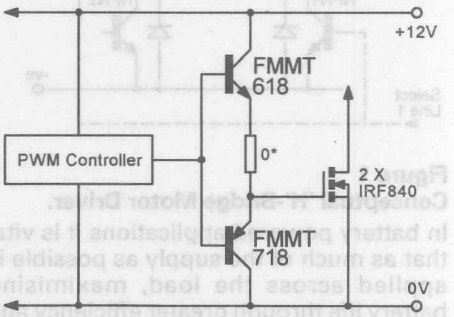


Figure 4
MOSFET/IGBT Complimentary Emitter Follower Gate Driver.

Where 5V logic provides a custom pulse width modulated controller, a buffer can be required to level translate, giving 10V or greater gate drive for the power switches. By using an FMMT2369 switching transistor, the circuit shown in Figure 5 converts 5V logic drive to a 12V gate drive signal. Driving the emitter of the FMMT2369 from the logic output avoids signal inversion.

The FMMT2369 gold doped switching transistor has a very short storage time which, combined with the high gain FMMT618 gives the circuit a turn-on time of only 20ns when driving a MOSFET with an effective input capacitance of 2nF. The FMMT718 helps make turn-off times even shorter, leading to reduced cross-conduction problems in bridge or push-pull converters.

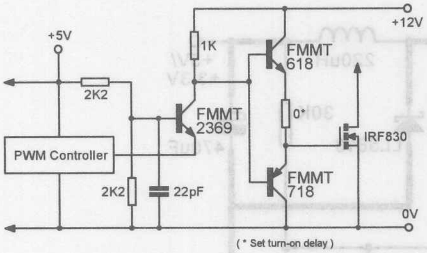


Figure 5
MOSFET/IGBT Gate Driver using Emitter Driven FMMT2369 for Buffering.

By giving excellent high current performance in a SOT23 package, the FMMT618 and FMMT718 replace SOT223 and SOT89 transistors in these gate drive circuits leading to cost and PCB area savings - particularly in very high frequency converters.

'H'-Bridge Siren Driver

Many modern burglar and automotive alarm sirens employ an 8Ω moving coil loudspeaker driven by a bipolar 'H'-bridge. Handling peak output currents of 2A, the TO126 or TO220 packaged output transistors normally used require parallel collector emitter diodes to divert destructive reverse transients generated by the inductive load. In the circuit shown in Figure 6, FMMT619 and FMMT720 SOT23 transistors replace these bulky and expensive leaded transistors, giving other savings too. High reverse h_{FE}, inherent in the matrix technology used to manufacture the Zetex transistors eliminates the need for parallel collector emitter protection diodes. The FMMT619 and FMMT720 conduct reverse collector current sufficiently well to clamp any inductive transients generated by the load.

A specially designed Zetex Siren Driver IC, the ZSD100, provides a variable frequency drive to the SOT23 'H'-bridge ensuring a very loud (and irritating!) noise is generated. The combination of the ZSD100 and a SOT23 'H'-bridge produces an extremely compact and inexpensive module.

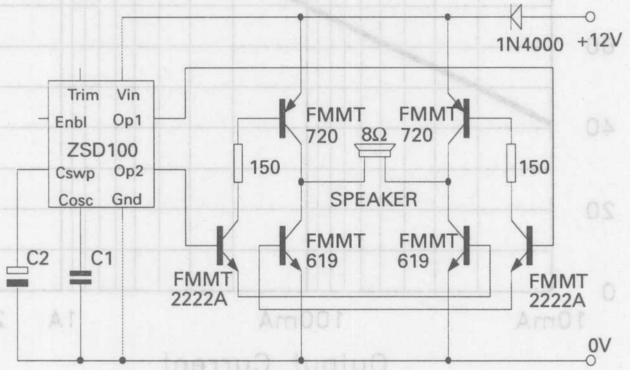


Figure 6
'H'-Bridge for Driving Moving Coil Loudspeakers within Alarm/Siren Systems.

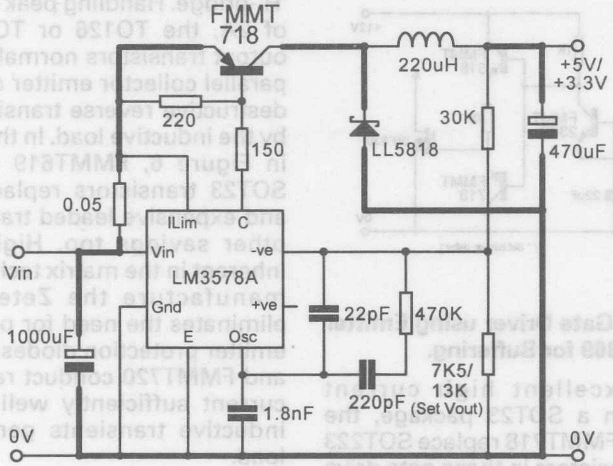


Figure 7
DC-DC Step-down Converter Effected with PNP SuperSOT. High Current Paths Shown in Bold.

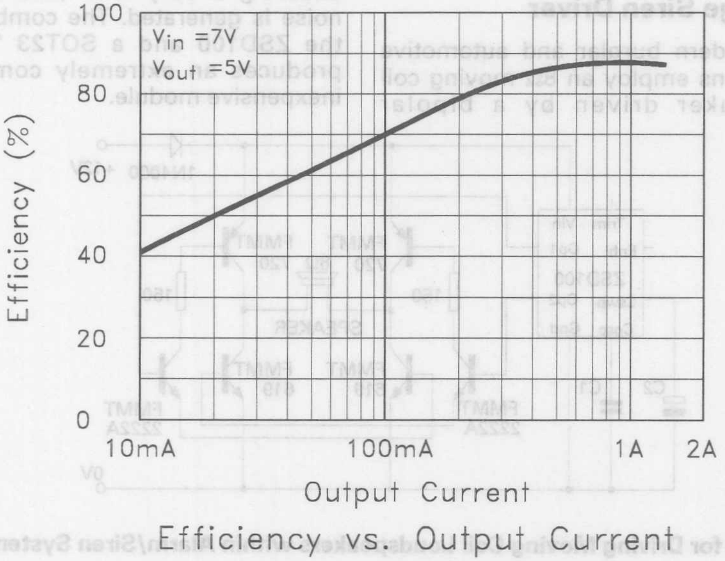


Figure 8
Efficiency Vs Output Current Profile for FMMT718 DC-DC Converter of Figure 7.

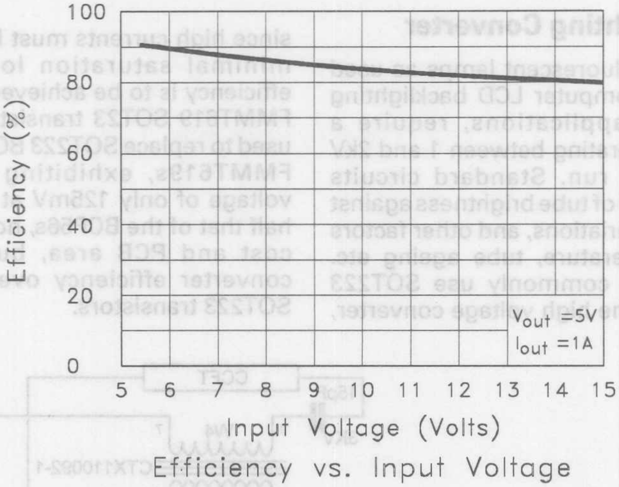


Figure 9
Efficiency Vs Input Voltage Profile for FMMT718 DC-DC Converter of Figure 7.

DC-DC Converter

Using standard PWM controllers it is easy to construct Buck step down converters with low component counts. Harder to achieve are designs that are both simple and efficient as required for today's battery operated equipment. The key to maximising efficiency is eliminating voltage drops in all high current areas.

In the Buck converter shown in Figure 7, the high current paths are via the 50mΩ sense resistor, the series switching transistor, output inductor L1 and the Schottky diode. Once the resistance of the output inductor has been minimised, the most critical component (particularly when V_{IN} approaches V_{OUT}) is the saturation voltage drop of the switching transistor. By using an FMMT718, which drops only 200mV @ 1.5A, this converter can operate at an efficiency of over 90% at minimum input voltage and an I_{OUT} of 1.5A. As the input voltage is increased, the operating gain of the switching transistor becomes more important. The high gain of this transistor allows base drive losses to be minimised leading to high efficiencies over a wide supply range. (Please refer to Figures 8 and 9).

The fast rise and fall times of the FMMT718 allows the converter to operate at 50kHz with minimal switching losses. At this frequency it is essential to use low ESR input and output capacitors, and keep any wires carrying switched high currents very short so as to minimise RFI and output ripple. (These wires are shown as bold in Figure 7).

The converter will operate from a supply of (V_{OUT} +0.5V) up to 16V. The converter's output voltage can be set to 5V or 3.3V by selection of R1. The circuit will supply loads from 0 to 1.5A, current limiting to around 2A with a shorted output.

LCD Backlighting Converter

Cold cathode fluorescent lamps as used for portable computer LCD backlighting and similar applications, require a converter generating between 1 and 2kV to strike and run. Standard circuits provide control of tube brightness against input supply variations, and other factors such as temperature, tube ageing etc. These circuits commonly use SOT223 transistors in the high voltage converter, since high currents must be passed with minimal saturation losses if good efficiency is to be achieved. In Figure 10, FMMT619 SOT23 transistors have been used to replace SOT223 BCP56 types. The FMMT619s, exhibiting a saturation voltage of only 125mV at 1A - less than half that of the BCP56s, not only reduces cost and PCB area, but also raises converter efficiency over the original SOT223 transistors.

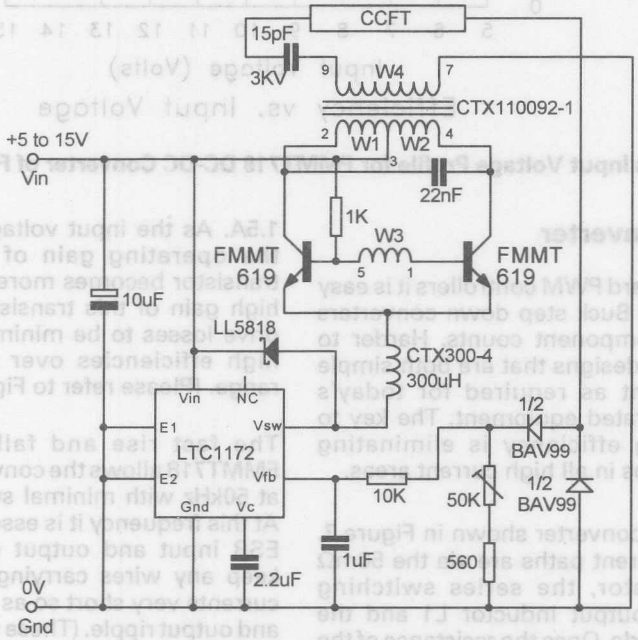


Figure 10
LCD Backlighting Inverter.

ZETEX

Application Note 13
Issue 2 January 1996

The ZCN0545A and ZCP0545A Low Power IGBTs

A Silicon/ Package Efficient Device for Compact Fluorescent Lamp (CFL) Ballasts and Static Switching

David Bradbury

IGBT Characteristics

The N-Type ZCN0545A and P-Type ZCP0545A are Insulated Gate Bipolar Transistors (IGBTs) available in the E-Line (TO92 compatible) package. The ZCN0545A is a 450V, 2A peak device and the ZCP0545A is a 450V, 0.8A peak device. With these devices Zetex has brought the advantages given by IGBTs to lower power and lower cost applications. (Please refer to Appendix A for summarised datasheet information).

IGBTs behave like a Darlington structure, utilising a MOSFET input device followed by a bipolar amplifier. The circuit symbols of the ZCN0545A and ZCP0545A are shown in Figure 1 and a simplified equivalent circuit for each is shown in Figure 2.

The most important characteristics of IGBTs are:-

1) MOSFET type **input resistance**, ideal for direct drive from microcontrollers.

2) **Low $R_{DS(on)}$**. For a given chip size and BV_{DSS} the $R_{DS(on)}$ of an IGBT is less than 10% of a standard high voltage MOSFET at high current.

3) Like a bipolar Darlington, the device needs a drain-source voltage of greater than 0.7V before current flows.

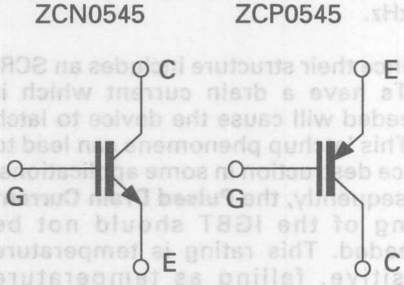


Figure 1
IGBT Symbols.

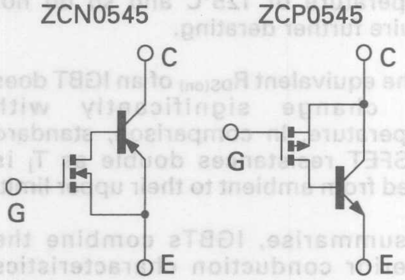


Figure 2
Conceptual Equivalent Circuit of the IGBT.

4) If the drain-source terminals of an IGBT are reverse biased, the drain-source diode of the input MOSFET cannot conduct since the base-emitter junction of the output bipolar transistor is in series. In many applications this can provide a useful **reverse blocking** capability.

5) **Switching speed** is dominated by the characteristics of the bipolar transistor which can be optimised for either speed or saturation voltage. The ZCN0545A and ZCP0545A are designed to be very fast, switching on in less than 20ns, and off in less than 300ns. This makes them suitable for switching applications up to 100kHz.

6) Since their structure includes an SCR, IGBTs have a drain current which if exceeded will cause the device to latch up. This latchup phenomena can lead to device destruction in some applications. Consequently, the **Pulsed Drain Current** rating of the IGBT should not be exceeded. This rating is temperature sensitive, falling as temperature increases. The data sheet specifications for the ZCN0545A and ZCP0545A are given for their maximum junction temperature of 125°C and so do not require further derating.

7) The equivalent $R_{DS(on)}$ of an IGBT does not change significantly with temperature. In comparison, standard MOSFET resistances double as T_j is raised from ambient to their upper limit.

To summarise, IGBTs combine the superior conduction characteristics given by bipolar Darlingtons with the ease of drive of MOSFETs. Though not quite as fast as MOSFETs, the ZCN0545A and ZCP0545A have been optimised to give very respectable switching speeds. In applications where the drive impedance is high, their low input capacitance can actually make them faster than equivalent current MOSFETs.

The output characteristics of the ZCN0545A and ZCP0545A are illustrated in Figures 3 and 4.

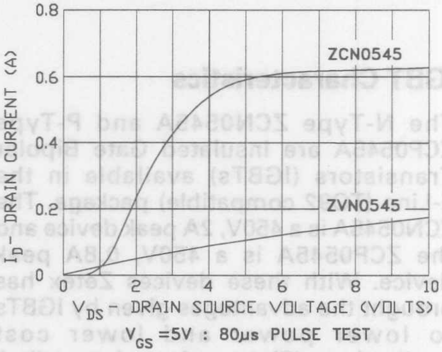


Figure 3
N-Type IGBT Vs N-Channel MOSFET saturation comparison.

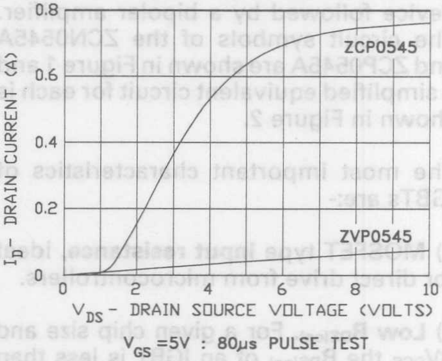


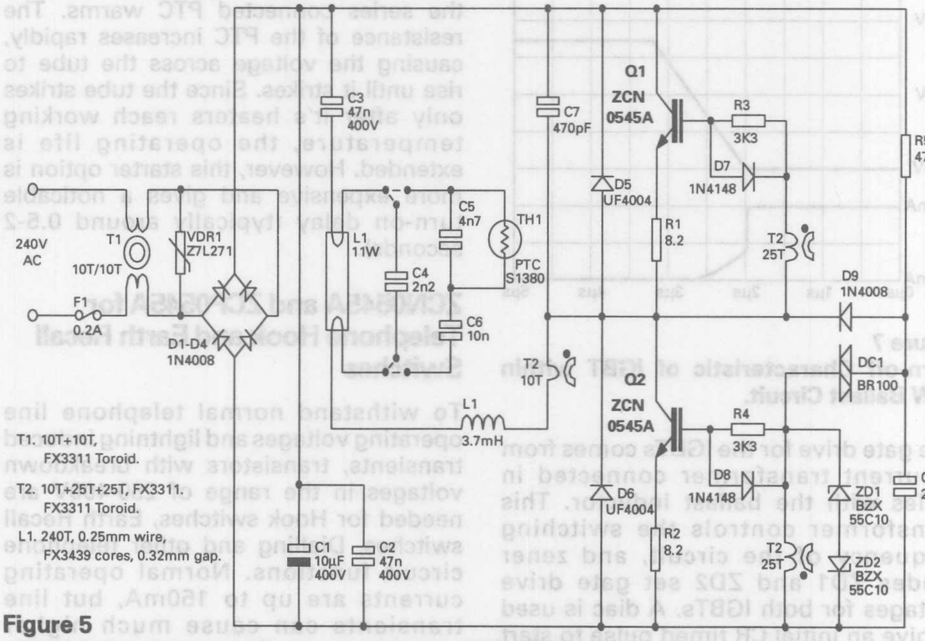
Figure 4
P-Type IGBT Vs P-Channel MOSFET saturation comparison.

They show typical I_D versus V_{DS} for a logic level gate drive of 5V. To indicate the improvement the IGBT structure gives over standard MOSFETs, a graph of the typical performance of an identical chip size MOSFET has been plotted for comparison.

ZCN0545A for Off-Line Fluorescent Lamp Ballast.

Figure 5 shows the circuit of an 11W off-line fluorescent lamp ballast using two ZCN0545A IGBTs. The efficiency of the circuit is such that it allows the E-Line IGBTs to replace the TO220 / TO126 bipolar or MOSFET transistors commonly used in this application. This both lowers component costs and gives a reduction in circuit size - critical in integral lamp/ballast designs.

The 300ns turn-off capability of the ZCN0545A would allow operation at up to 100kHz but the working frequency of the design was set at 40kHz to minimise losses and HF interference. By controlling the phase of the current flowing in the IGBTs so that cross-conduction does not occur, switching losses have been virtually eliminated. Also, the low effective $R_{DS(on)}$ of the ZCN0545A keeps conduction losses to around 60mW in each device. Figure 6 shows the voltage and current waveforms of the IGBTs. Figure 7 shows an expanded view of the critical turn-off behaviour of the ZCN0545A. Note in particular that the drain current falls to zero before the drain voltage rises significantly, ensuring low switching losses.



T1. 10T+10T,
FX3311 Toroid.
T2. 10T+25T+25T, FX3311.
FX3311 Toroid.
L1. 240T, 0.25mm wire,
FX3439 cores, 0.32mm gap.

Figure 5
11W Off-line Compact Fluorescent Lamp (CFL) Ballast.

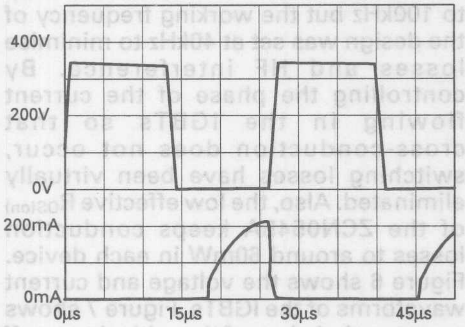


Figure 6
IGBT Voltage and Current Waveforms within 11W ballast Circuit.

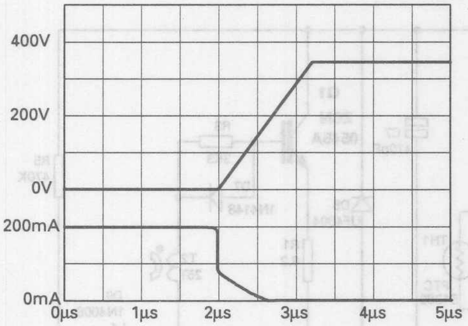


Figure 7
Turn-off Characteristic of IGBT within 11W Ballast Circuit.

The gate drive for the IGBTs comes from a current transformer connected in series with the ballast inductor. This transformer controls the switching frequency of the circuit, and zener diodes ZD1 and ZD2 set gate drive voltages for both IGBTs. A diac is used to give an initial CR timed pulse to start the circuit, and transformer T1 and voltage dependent resistor VDR1 control line borne transients and interference.

Two strike circuits can be used. The simplest (and lowest cost) is to use a single capacitor which gives the circuit an instant start characteristic. However, this has the disadvantage that the lamp strikes before the heaters warm fully, leading to tube end blackening and some reduction of tube life when the unit is switched on and off frequently.

An improved mode of operation comes from using a two capacitor/PTC starter combination. At turn-on, a 10nF capacitor forces a high heater current to flow until the series connected PTC warms. The resistance of the PTC increases rapidly, causing the voltage across the tube to rise until it strikes. Since the tube strikes only after it's heaters reach working temperature, the operating life is extended. However, this starter option is more expensive and gives a noticable turn-on delay (typically around 0.5-2 seconds).

ZCN0545A and ZCP0545A for Telephone Hook and Earth Recall Switches

To withstand normal telephone line operating voltages and lightning induced transients, transistors with breakdown voltages in the range of 250-400V are needed for Hook switches, Earth Recall switches, Dialling and other telephone circuit functions. Normal operating currents are up to 150mA, but line transients can cause much higher currents.

Figure 8 shows the line interface section of a feature phone. In this circuit Q1, a ZCN0545A, is being used as an earth recall switch. Under the direction of a function controller this transistor grounds the negative supply rail of the telephone circuit, requesting a special switchboard service. Normally, N-channel MOSFETs are used for earth recall switches since the phone power supply is shorted during recall operation so low drive power is mandatory. The ZCN0545A IGBT however provides the same high input impedance as a MOSFET, but also gives lower on-voltage drops at the high supply currents that can occur with short telephone lines.

Q2, a ZCP0545A, is used as an electronic hook switch as needed in some feature phones. It's gate drive is controlled by a ZTX457 bipolar transistor wired as a current source to limit drain current when the circuit is operating during a line transient. The P-type hook switch function is normally achieved using an expensive P-channel MOSFET or a PNP/NPN quasi-Darlington arrangement, but the ZCP0545A IGBT provides a lower cost and or reduced component count solution for this application.

To avoid ring-trip problems, the transient suppressors used to protect telephone handsets must not conduct for input voltages below 200V. Tolerancing on these components leads to the selection of suppressors that under worst case conditions will allow peak voltages of over 270V during a transient. The 450V BV_{DSS} of the ZCN0545A and ZCP0545A eases the design of the transient protection circuit, normally a tight balance between ring-trip problems and adequate

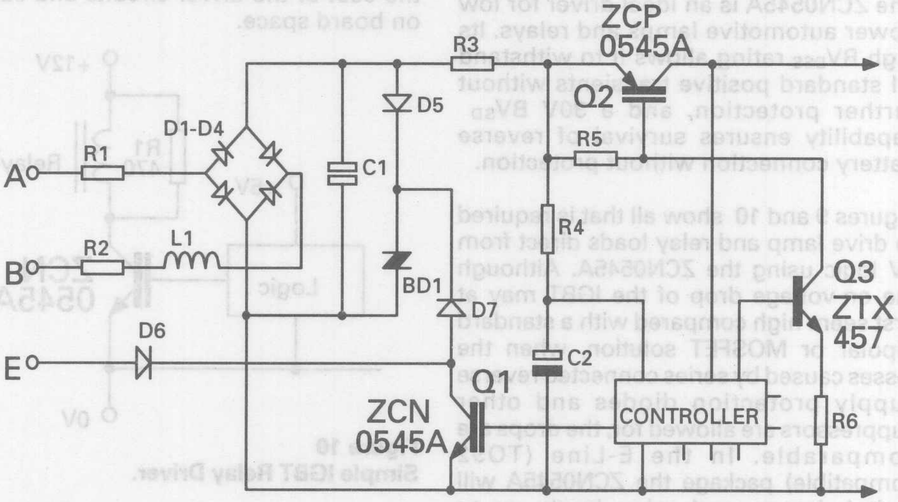


Figure 8
Telephone Line Interface Circuit.

protection. Transient protection for the design in Figure 8 is provided by D5 and BD1.

ZCN0545A for Automotive Lamp and Relay Drivers.

Automotive lamp and relay drivers have to withstand extreme operating conditions. Behind an instrument panel temperatures can vary from a low of -40°C to over +85°C. Such drivers may have to withstand reverse battery connection, 24V rapid battery charging, a 100-240V 0.5J +ve transient, an 80V 50J load dump transient, and a number of other lower energy supply transients. The driver device will pass many times the normal load current if enabled during a supply transient. Consequently few driver transistors can be used without extensive transient and reverse supply protection.

The ZCN0545A is an ideal driver for low power automotive lamps and relays. Its high BV_{DSS} rating allows it to withstand all standard positive transients without further protection, and a 30V BV_{SD} capability ensures survival of reverse battery connection without protection.

Figures 9 and 10 show all that is required to drive lamp and relay loads direct from 5V logic using the ZCN0545A. Although the on-voltage drop of the IGBT may at first seem high compared with a standard bipolar or MOSFET solution, when the losses caused by series connected reverse supply protection diodes and other suppressors are allowed for, the drops are comparable. In the E-Line (TO92 compatible) package the ZCN0545A will switch lamps and relay loads up to 180mA.

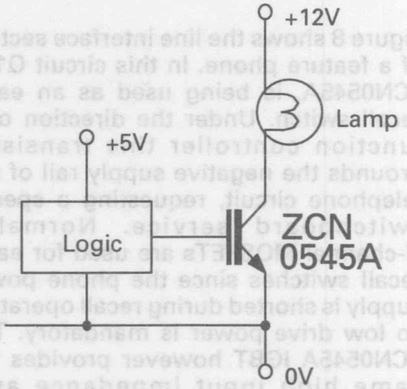


Figure 9
Simple IGBT Bulb Driver.

In Figure 10, a high value low power resistor is connected across the relay coil to suppress transients generated by the relay during turn-off. Apart from this, no protection components are required for either circuit. This greatly reduces the cost of the driver circuits and saves on board space.

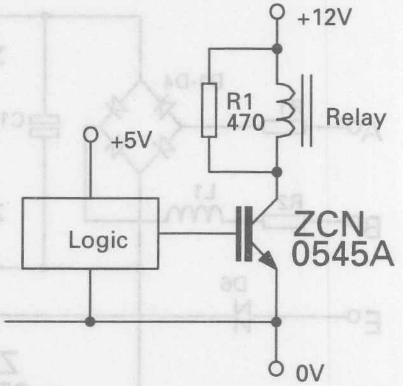


Figure 10
Simple IGBT Relay Driver.

Appendix A

ZCN0545A and ZCP0545A Summary Data-Sheet

ABSOLUTE MAXIMUM RATINGS

PARAMETER		SYMBOL	ZCN0545A	ZCP0545A	UNIT
Forward Drain-Source Voltage		V_{DS}	450	-450	V
Reverse Drain Source Voltage		V_{SD}	30	-20	V
Continuous Drain Current		I_D	0.32	-0.32	A
Pulsed Drain Current	@ T_{amb}=25°C	I_{DMR}	2	-0.8	A
	@ T_{amb}=125°C	I_{DM}	1	-0.4	A
Gate-Source Voltage		V_{GS}	±20		V
Power Dissipation		P_{tot}	0.6		W
Operating and Storage Temperature Range		T_j:T_{stg}	-55 to +125		°C

ELECTRICAL CHARACTERISTICS (at T_{amb} = 25°C unless otherwise stated)

PARAMETER	SYMBOL	ZCN0454A MIN.	ZCN0454A MAX.	ZCP0545A MIN.	ZCP0545A MAX.	UNIT	CONDITIONS.
Forward Drain-Source Breakdown Voltage	BV_{DSS}	450		-450		V	V_{GS}=0V
Reverse Drain-Source Breakdown Voltage (4)	BV_{SD}	30		-20		V	I_D=1mA
Gate-Source Threshold Voltage	$V_{GS(th)}$	1	3	-1	-3.5	V	I_D=1mA, V_{DS}= V_{GS}
Zero Gate Voltage Drain Current	I_{DSS}		10 400		-20 -2000	µA µA	V_{DS}=450V, V_{GS}=0V V_{DS}=360V, V_{GS}=0V, T=125°C (2)
Drain Source Saturation Voltage (1)	$V_{DS(SAT)}$		3 3		-3 -3	V V	I_D=500mA, V_{DS}=10 V I_D=250mA, V_{DS}=5 V
Static Drain-Source On-State Resistance (1)	$R_{DS(on)}$		6		6	Ω	V_{GS}=10V,I_D=0.5A
Input Capacitance (2)	C_{iss}		90		120	pF	V_{DD}≈25V, V_{GS}=0V, f=1MHz
Switching Times (2)(3)	t_{on}		150		150	ns	V_{DD}≈25V, V_{GEN}=10V I_D=1A, R_{GS}=50Ω
	t_{off}		300		350	ns	

1) Measured under pulsed conditions. Pulse width=300µs. Duty cycle ≤2%
(2) Sample test.
(3) Switching times measured with 50Ω source impedance and <5ns rise time on a pulse generator.
(4) One minute maximum duration. Exceeds common international automotive reverse battery test specifications.

Please refer to datasheets for full device characterisation.

Transistor Considerations for LCD Backlighting

High Efficiency DC to AC Conversion

Neil Chadderton

Introduction

LCD Backlighting has generated widespread interest from many diverse disciplines within the engineering industry. This has no doubt been fueled by the trend to portability and particularly to the enormous growth of the computing market. Products such as notebook, laptop, and palmtop personal computers, portable televisions, viewcams, point of sale terminals, automotive dashboards, avionics displays, metering and instrumentation usually employ an LCD screen, and as such require a means of backlighting. To date the most prevalent method has been to use a small cold cathode fluorescent (CCFL) tube that is usually integrated with a reflector/diffuser into the display unit. The CCFL power consumption can account for a significant portion (up to 50%) of the total system requirement. Therefore to achieve marketable advantages in battery life and re-charge frequency, much attention must be applied to the CCFL power supply, so as to attain the highest possible conversion efficiency.

This problem has been the focus of many electronic component vendors: much research and design effort being invested in order to offer system designers the most attractive components/solutions in terms of efficiency, cost, weight, and size. Many of the analog IC companies have published application specific reports, and characterised or developed specifically, integrated circuits for the application.

This note acknowledges this work, and will draw upon such sources and reproduce these vendors' circuits where appropriate (a list of references is included in Appendix A) but it is focused primarily on the transistor requirements – their mode of operation within the backlighting circuit, important parameters, and their impact on the system efficiency.

CCFL Lamp Characteristics

An understanding of the requirements for the backlighting power supply should begin with a description of the load involved. The fluorescent tube presents a serious challenge to the circuit designer. Around 1kV is required to strike the tube (initiate conduction), at which event the tube's gaseous contents ionise and it begins to conduct at a lower sustaining voltage – thus a negative resistance characteristic is evident. Other power supply constraints include an intolerance of DC current, a sensitivity to waveform crest factor, and RFI criteria.

The curve tracer plots shown in Figures 1 and 2 show the negative resistance region for two typical CCFL units: – the first for a 150mm linear, 10mm diameter backlight tube for a laptop display, and the other a "U" tube as produced for a car dashboard display. Referring to figure 1, the high striking voltage can be seen at 560V and the negative resistance excursion to 240V is self evident. Similarly, these values for Figure 2 are 1240V and 900V. Note should also be made of the slope impedance in the conducting state. The power supply must accommodate this, and in some cases provision made to regulate the lamp current to ensure a long tube life.

For drive waveforms at low frequencies, a fluorescent tube has time to react to the changing waveform potential, and effectively re-strikes on each reversal of the waveform polarity, (perceived as flicker on line frequency units). At high drive waveform frequencies, this effect is not apparent, and the lamp can be approximated to a resistive load. Usual operating frequencies range from 25 to 120kHz, this being dictated by consideration of inaudibility requirements, converter inductor size, and at the extreme, parasitic and HV-lead-to-ground coupling capacitance.

Basic Operation Of Converter

The drive requirements dictated by the CCFL tube's behaviour and preferred operating conditions can be achieved by the resonant push-pull converter shown in Figure 3. This is also referred to as the Royer Converter, after G.H. Royer who proposed the topology in 1954 as a power converter. (Note: Strictly speaking the backlighting converter uses a modified version of the Royer converter – the original used a saturating transformer to define the operating frequency, and therefore produced a squarewave drive

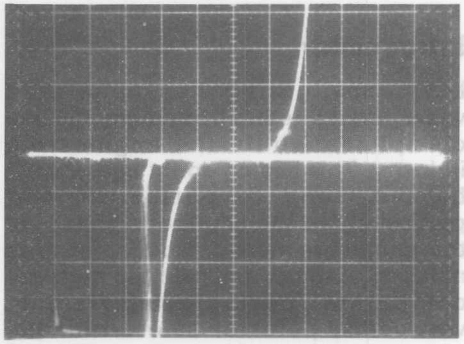


Figure 1.
CCFL Characteristics - 150mm linear; 100V/div horizontal, 200μA/div vertical.

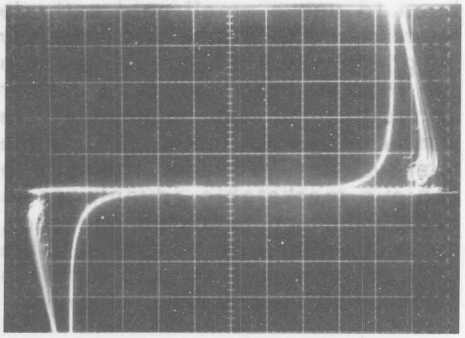


Figure 2.
CCFL Characteristics - "U" tube; 200V/div horizontal, 1mA/div vertical.

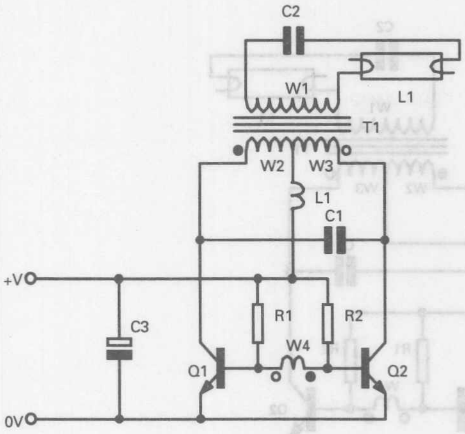


Figure 3.
Generalised Royer Converter.

waveform). The circuit looks simple but this is very deceptive: many components interact, and while the circuit is capable of operation with widely varying component values, (useful during development) optimisation is required for each design to achieve the highest possible efficiencies.

Transistors Q1 and Q2 are alternatively saturated by the base drive provided by the feedback winding W4. The base current is defined by resistors R1 and R2. Supply inductor L1 and primary capacitance C1 force the circuit to run sinusoidally thereby minimising harmonic generation and RFI, and providing the preferred drive waveform to the load. Voltage step-up is achieved by the W1:(W2 + W3) turns ratio. C2 is the secondary winding ballast capacitor, and effectively sets the tube current.

Prior to the tube striking, or when no tube is connected, the operating frequency is set by the resonant parallel circuit comprising the primary capacitance C1, and the transformer's primary winding W2+W3. Once the tube has struck, the ballast capacitor C2 plus distributed tube and parasitic capacitances are reflected back through the transformer, and the operating frequency is lowered.

The secondary load can become dominant in circuits with a high transformer turns ratio, Eg. those designed to operate from very low DC input voltages.

Each transistor's collector is subject to a voltage= 2 x $\pi/2$ x V_S, (or just π x V_S) where V_S is the DC input voltage to the converter. (The $\pi/2$ factor being due to the relationship between average and peak values for a sinewave, and the x2 multiplier being due to the 2:1 autotransformer action of the transformer's centre-tapped primary). This primary voltage is stepped up by the transformer turns ratio Ns:Np, to a high enough level to reliably strike the tube under all conditions:- starting voltage is dependent on display housing, location of ground planes, tube age, and ambient temperature.

The basic converter shown in Figure 3 is a valid and useful circuit that has been utilised for many systems and indeed offered as a sub-system by several manufacturers.

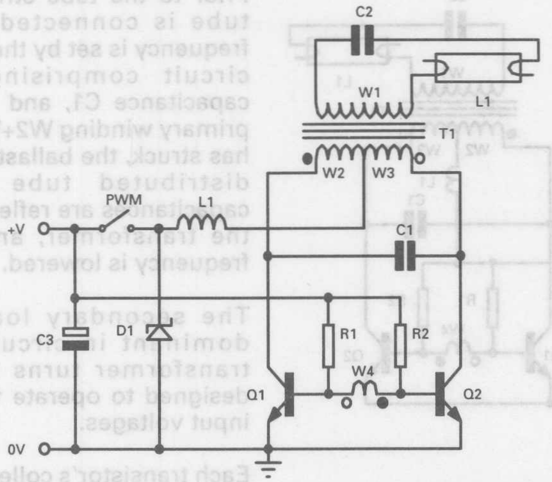


Figure 4a.
Royer Converter With PWM Control - High Side Current Fed Version.

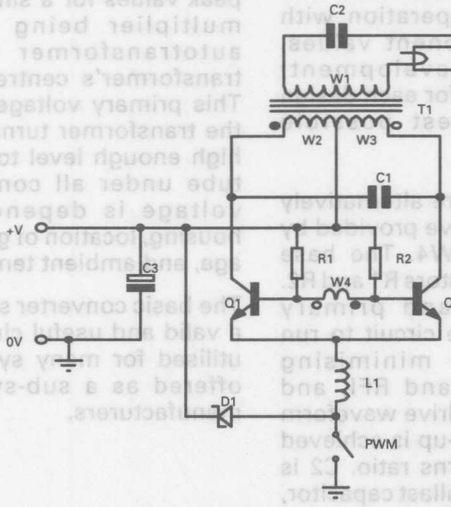


Figure 4b.
Royer Converter With PWM Control - Low Side (or tail) Current Fed Version.

Backlight Converters Within Control Loops

Variations on the basic topology are possible, perhaps the most important being to include the converter within a control loop. This can be used to regulate the tube current:- this maximises tube lifetime, ensures a constant light output as the battery pack voltage decreases, and enables adjustment of tube brightness. The usual form of the circuit is to employ a Buck or step-down converter (directly from the battery pack to increase efficiency) feeding the centre tap of the transformer, or the emitter current of the transistors, depending on the controller's technology and capability. Figures 4a and 4b show these arrangements in conceptual form. The controller can monitor the tube current directly in the secondary, or in some recent systems, by the primary current. This latter method allows the tube to be fully floating thus minimising HV losses.

Figure 5 shows a circuit published by linear IC manufacturer LINEAR TECHNOLOGY CORP. that exhibits a significant efficiency improvement over previous designs; primarily due to the choice of the **ZETEX FZT849**. It is based on the Buck converter current fed Royer scheme of Figure 4b, and monitors the lamp's current directly by averaging the positive half cycles of lamp current, and applying this signal to the controller's feedback pin. The electrical conversion efficiency using this form of circuit can be very high, the stated value for Figure 5 being 88%. Higher efficiencies up to 92% are possible by using larger transformers to reduce copper and core losses.

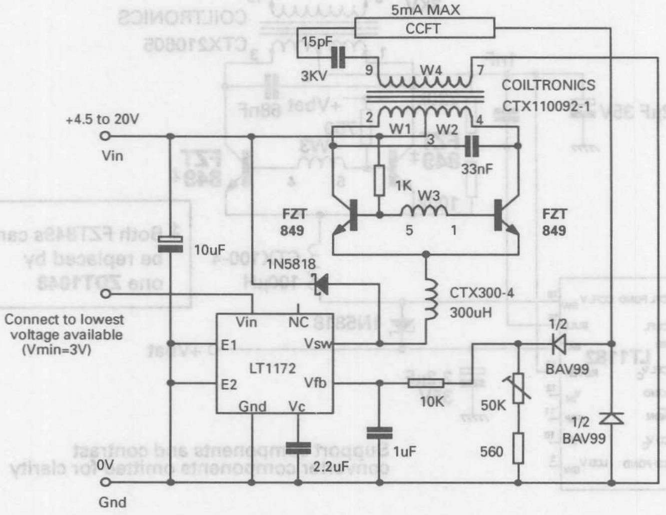


Figure 5.
Linear Technology LCD Backlight Converter.

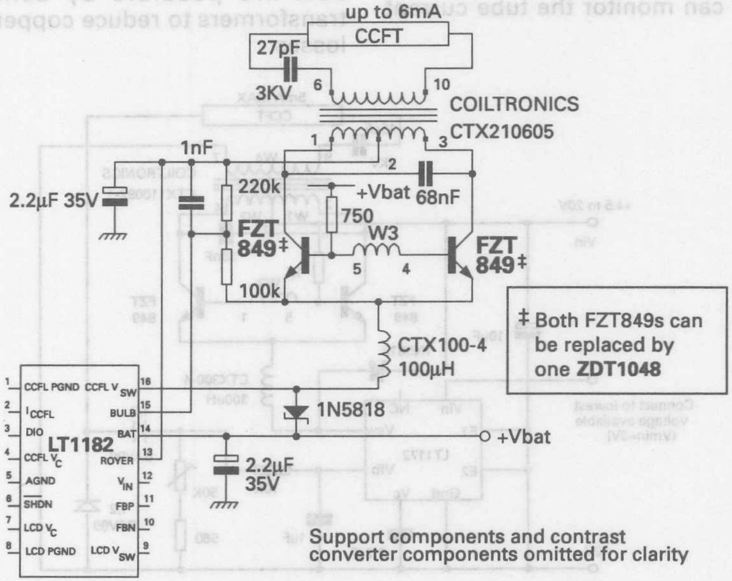
Figure 6 shows Linear Technology's latest design using the LT1182 and the Zetex ZDT1048 dual transistor. The LT1182 provides a low component count circuit and contains all control functions for the Royer converter, and the control/switch for the LCD contrast converter within one package. Primary Royer converter current is sensed by the IC, so that the CCFL tube can be operated in a "floating" mode thereby decreasing losses in the secondary circuit. The FZT849 transistors, or the ZDT1048 dual package are preferred options for this converter circuit.

Detailed reports on these circuits can be found via the references listed.

Figure 7 shows an oscillograph of the transistor's operating conditions in such a circuit. The Collector-Emitter voltage peaks at 28V (less than π x V_S due to the lamp load); the Emitter current is almost constant at 0.5A (with a ripple component dependent on the Buck inductor); and the base voltage appears as a clipped (due the transistor's V_{BE}) version of the primary waveform.

Requisite Transistor Characteristics

The relatively low operating frequency as required by the backlighting Royer Converter (to minimise HV parasitic capacitance losses), and the ease of transformer drive, makes this circuit particularly suitable for bipolar transistor



‡ Both FZT849s can be replaced by one **ZDT1048**

Support components and contrast converter components omitted for clarity

Figure 6.
Linear Technology Floating Tube LCD Backlight Converter.

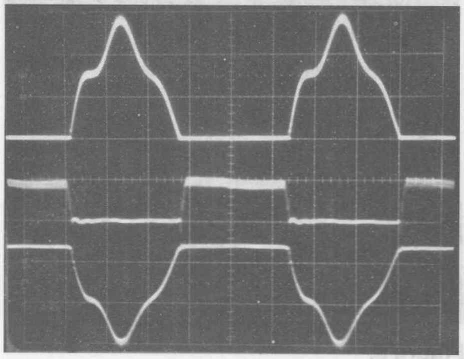


Figure 7.
Royer Converter Operating Waveforms: V_{CE} 10V/div; I_E 0.5A/div; V_{BE} 2V/div respectively, 2µs/div horizontal .

implementation. This isn't to exclude MOSFET based designs (some IC vendors have specified MOS as this suits their technology) but in terms of equivalent on-resistance and silicon efficiency, the low voltage bipolar device has no equal. For example, the **ZETEX ZTX849** E-Line (TO-92 compatible) transistor exhibits a $R_{CE(sat)}$ of 36mΩ. This can only be matched by a much larger (and expensive) MOSFET die, only available in TO-220, D-Pak, and similar larger packages.

The important transistor characteristics are voltage rating, $V_{CE(sat)}$, and h_{FE}, and are detailed below.

The voltage rating required deserves some thought with respect to the standard transistor breakdown parameters, as it is possible to over-specify a device on grounds of voltage rating, and thereby incur a reduction in efficiency due to unnecessary on-resistance losses. The primary breakdown voltage BV_{CBO}, of a planar bipolar transistor depends on the epitaxial layer - specifically its thickness and resistivity. The breakdown voltage of most interest to the designer is usually that attained across the Collector-Emitter (C-E) terminals. This value can vary between the primary breakdown BV_{CBO} and a much lower voltage dependent on the state of the base terminal bias.

[The breakdown mechanism is caused by the avalanche multiplication effect, whereby free electrons can be imparted with sufficient energy by the reverse bias electric field such that any collisions can lead to ionisation of the lattice atoms. The free electrons thus generated are then accelerated by the field and produce further ionisation. This multiplication of free carriers increases the reverse current dramatically, and so the junction effectively clamps the applied voltage. The base terminal can obviously influence the junction current –thereby modulating the voltage required for a breakdown condition.]

Figure 8 shows how the breakdown characteristic is seen to vary for different circuit conditions. The BV_{CEO} rating (or when the base is open circuit) allows the Collector-Base (C-B) leakage current I_{CBO} to be effectively amplified by the transistor's β thus significantly increasing the leakage component to I_{CEO}. Shorting the Base to the Emitter (BV_{CES}) provides a parallel path for the C-B leakage, and so the voltage required for breakdown is higher than the open base condition. BV_{CER} denotes the case between the open and shorted base options:- R indicating an external base-emitter resistance, the value of which is typically 100 to 10kΩ. BV_{CEV} or BV_{CEX} is a special case where the base-emitter is reverse biased; this can provide a better path for the C-B leakage, and so this rating yields a voltage close to, or coincident with the BV_{CBO} value.

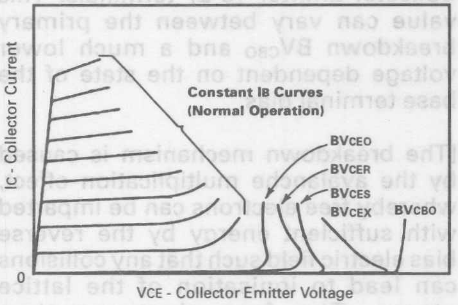


Figure 8.
Voltage Breakdown Modes of Bipolar Transistor.

Figure 9 shows a curve tracer view of the relevant breakdown modes of the **ZTX849** transistor, including a curve showing the device in the "on" state. Curves 1 and 2 are virtually coincident and show BV_{CBO} and BV_{CES} respectively. Curve 3 shows the BV_{CEV} case with an applied base bias (V_{EB}) of -1V. Curve 4 shows BV_{CEO} at approximately 36V. Curve 5 is a BV_{CE} curve, showing how the breakdown condition is affected by a positive base bias of 0.5V.

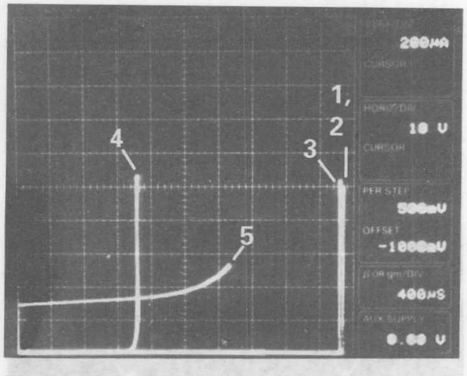


Figure 9.
Breakdown modes of the ZTX849 Bipolar Transistor.

The BV_{CEV} rating has particular relevance to the Royer Converter, as can be surmised from Figure 7. Examination of this will show that the transistor only experiences the high C-E voltage when the base voltage has been taken negative by the feedback winding, these events of course being in perfect synchronism. An expanded view of the C-E and B-E waveforms is shown in Figure 10.

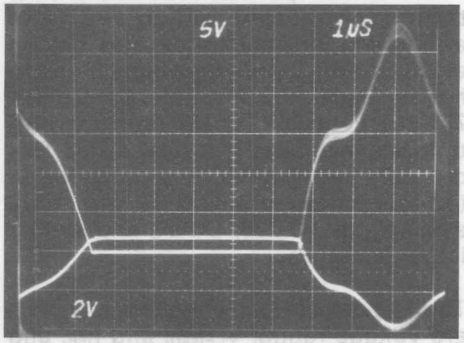


Figure 10.
Royer Converter: V_{CE} and V_{BE} Waveforms 5V/ div and 2V/ div respectively.

[Note: The voltage applied by the feedback winding must not exceed the BV_{EBO} of the transistor. This is specified at 5V usually, against an actual of 7.5 to 8.5V].

The $V_{CE(sat)}$ and h_{FE} parameters have a direct bearing on the circuit's electrical conversion efficiency. This is especially true of low voltage battery powered systems, due to the high current levels involved. Selection of standard LF amplifier transistors provides far from ideal results; these parts are for general purpose linear and non-critical switching use only. The high $V_{CE(sat)}$ inherent to these parts, and low current gain could reduce circuit efficiency to less than 50%. For example, the stated $V_{CE(sat)}$ maximum measured at 500mA, for the **FZT849** SOT223 transistor, and a LF device sometimes quoted as a suitable Royer Converter transistor are 50mV and 0.5V respectively. Eg.

	$V_{CE(sat)}$	$@I_C$	I_B
FZT849	50mV	0.5A	20mA
BCP56	0.5V	0.5A	50mA

To address the $V_{CE(sat)}$ issue, large power transistors are occasionally specified. Unfortunately their capacitance, and characteristic low base transport factor (a feature of Epitaxial Base devices) can lead to problems with cross-conduction losses due to long storage and switching times. The current gain is also important, as the losses in the base bias can be significant to the overall figure; judicious selection of the bias resistor to ensure a minimum $V_{CE(sat)}$ while preventing base overdrive needs to consider supply variation, maximum lamp current, and transistor h_{FE} minimum value and range.

For the above reasons, transistors designed and optimised for high current switching applications offer the most cost-effective and efficient solutions. The table presented in Appendix C lists several **ZETEX** transistors that are eminently suitable for the Royer converter. All of these parts offer outstanding $V_{CE(sat)}$ and high current performance for their size, and many are so-called "Super-β" transistors; thereby helping to simplify and improve drive current requirements. Figure 11 shows the $V_{CE(sat)}$ exhibited by the ZTX1048A for a range of forced gain values. This device is one of the ZTX1050 series of transistors that employ a scaled up variant of the highly efficient Matrix geometry, developed for the ZETEX "SuperSOT" series. This enables a $V_{CE(sat)}$ performance similar to the ZTX850 series at the low to moderate currents relevant to this application, though utilising a smaller die, and therefore providing a cost and possibly a space saving advantage.

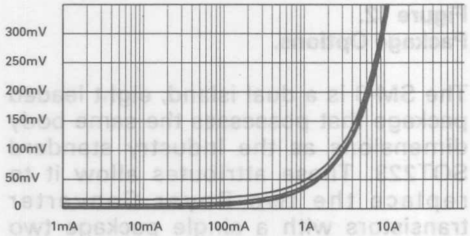


Figure 11.
$V_{CE(sat)}$ v I_C for the ZTX1048A Bipolar Transistor: Forced gains of 10,20,50,100.

Package Options

ZETEX can offer a range of packages to allow complete circuit size and layout optimisation. Figure 12 illustrates these, from the TO92 compatible E-Line through-hole package, to surface mount options SOT23, SOT223, and SM-8.

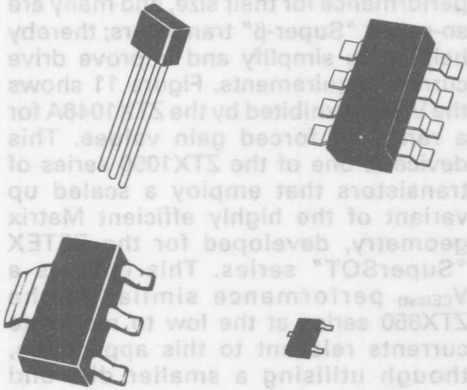


Figure 12.
Package Options.

The SM-8 is a dual island, eight leaded package that possesses the same body dimensions as the industry standard SOT223. These attributes allow it to replace the two Royer Converter transistors with a single package two chip device, yielding a significant cost and space saving.

For example, the '1048A transistor is available as an uncommitted dual within the SM8 package as the **ZDT1048**.

Conclusions

The advanced transistor geometries, and optimised processing employed by **ZETEX** leads to a range of transistors that are ideally suited to the LCD backlighting inverter application. Attention has been applied to specifying a range of devices relevant to, and exhibiting a superior performance within the Royer inverter topology.

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Applications Note 55 August 1993
Jim Williams
Linear Technology Corp.,

"A Fourth Generation of LCD Backlight Technology - Component and Measurement Improvements Refine Performance" Application Note 65 October 1995
Jim Williams
Linear Technology Corp.

"Switching and Linear Power Supply, Power Converter Design"
A. Pressman
Hayden Press.

Appendix A

LT1070, 1170 Series Switching Regulators
LT1182, 1183 CCFL/LCD Contrast Dual Switching Regulator
Linear Technology Corporation,
1630 McCarthy Blvd.,
Milpitas, CA 95035-7487
TEL: (408) 432 1900

Linear Technology (UK) Ltd.,
TEL:(01276) 677676

Linear Technology KK
Tokyo, 102 JAPAN
TEL: 81-3-3237-7891

Appendix B

CCFL Inverter Transformer and Inductor Manufacturers

Coiltronics Inc.,
TEL: (407) 241-7876
(Transformers and inductors)
Represented by METL in the UK
TEL: 01844-278781

Sumida Electric Co., Ltd.
Tokyo 125 JAPAN
TEL: 03-3607-5111
(Inductors)
Represented by ACAL Electronics Ltd.,in the UK
TEL: 0344-727272

Sumida Electric (USA) Co., Ltd
TEL: (708) 956-0666
(Transformers and Inductors)

Coilcraft
TEL: (708) 639-6400
(Inductors)

Coilcraft (UK)
TEL: 0181-301-3553

Newport Components Ltd.,
TEL: 01908-615232
(Inductors)

Pico Electronics Inc.,
NY 10552
TEL: (914) 699-5514
(Inductors)
Represented by Ginsbury Electronics Ltd., in the UK
TEL: 01634-290040

Appendix C

ZETEX Royer Converter Transistors

Device	BV_{CEV} * V	BV_{CES} / BV_{CBO} V	B_{VEBO} V	I_C (DC) A	h_{FE}	@ I_C / V_{ce} A / V	$V_{CE(sat)}$ V	@ I_C / I_B A / A	Package	Surface Mount Option
ZTX849	–	80	6	5	100 - 300	1 / 1	25mV typ 50mV Max	0.5 / 0.02	E-Line	FZT849 (SOT223)
ZTX869	–	60	6	5	300 min	1 / 1	20mV typ 50mV Max	0.5 / 0.01	E-Line	FZT869 (SOT223)
ZTX689B	–	50(typ)	5	3	450 min	1 / 2	60mV typ	0.5 / 0.005	E-Line	FZT689B (SOT223)
FMMT619 (SuperSOT)	–	50	5	2	200 min	1 / 2	55mV typ 125mV typ 200mV Max	0.5 / 0.01 1.0 / 0.01	SOT23	-
ZTX1048A	50	50	5	4	300 - 1200	1 / 2	24mV typ 45mV Max	0.5 / 0.02	E-Line	ZDT1048 (SM-8)
ZTX1049A	80	80	5	4	300 - 1200	1 / 2	35mV typ 60mV Max	0.5 / 0.02	E-Line	ZDT1049 (SM-8)

* If specified. For those devices that don't include a BV_{CEV} test, the actual value will be close to the BV_{CES}/BV_{CBO} figure - please refer to text.

ZETEX

Application Note 15
Issue 1 November 1995

Features and Applications of the FMMT617 and FMMT717 "SuperSOT" SOT23 Transistors

3A NPN and 2.5A PNP SOT23 Bipolar Devices

David Bradbury
Neil Chadderton

Introduction

The following note describes some of the features, benefits and applications of the FMMT617 NPN and FMMT717 PNP SOT23 transistors developed by Zetex. These devices extend the FMMT620 and FMMT720 high performance surface mount bipolar ranges. Specially optimised for the stringent requirements of battery powered systems, these tiny SOT23 packaged devices will replace much larger bipolar and MOSFET transistors, leading to significant savings in component costs and PCB sizes. Indeed the FMMT617 and FMMT717 transistors outperform many SOT223 and SOT89 types plus all SOT23 transistors presently available world-wide in terms of current handling and low losses.

Features

As can be seen in Table 1, the FMMT617 is a 15V NPN transistor capable of switching loads of up to 3A continuous, 12A peak. Designed to give a high mid-band gain of 450, the matrix chip geometry ensures this level of

Parameter	FMMT617	FMMT717	Units
Polarity	NPN	PNP	
BV_{CEO}	15	12	V
$I_{C\ continuous}$	3	2.5	A
$I_{C\ maximum}$	12	10	A
Mid-band h_{FE}	450	450	
Typical h_{FE}	320	275	
$@I_C$	3	2.5	A
Typical $V_{ce(sat)}$	150	160	mV
$@I_C$	3	2.5	A
Ptot	625	625	mW

Table 1
Parametric Overview.

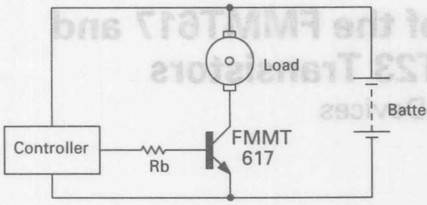


Figure 1
Generic Load Driver For Text Discussion.

performance is maintained well above the 3A rating. For example, it's typical h_{FE} at 5A is still around 240 and even at 12A the h_{FE} is a creditable 80, thus allowing the devices to handle very high current pulses, and tolerate switching transients. Giving a saturation voltage of only 150mV at 3A for a forced gain of 60, the FMMT617 is a highly efficient switch.

The FMMT717, though not quite as good as it's NPN counterpart, still gives excellent performance. It is a 12V PNP device rated at a collector current of 2.5A continuous, 10A peak. Also designed to give a mid-band gain of 450, the h_{FE} is still around 275 at 2.5A I_c. The saturation voltage of this part is a low 160mV at 2.5A.

The key features vital in battery powered equipment are low $V_{CE(sat)}$ and high h_{FE}. Both these parameters are important determinants of losses and hence battery life. Consider the simple motor driver circuit using the FMMT617 shown in Figure 1.

Conduction losses and the much lower base current losses occurring in the FMMT617 are charted in Figure 2. Conduction losses are given by $V_{CE(sat)} \times I_c$ and base current losses by $I_b \times V_{be}$.

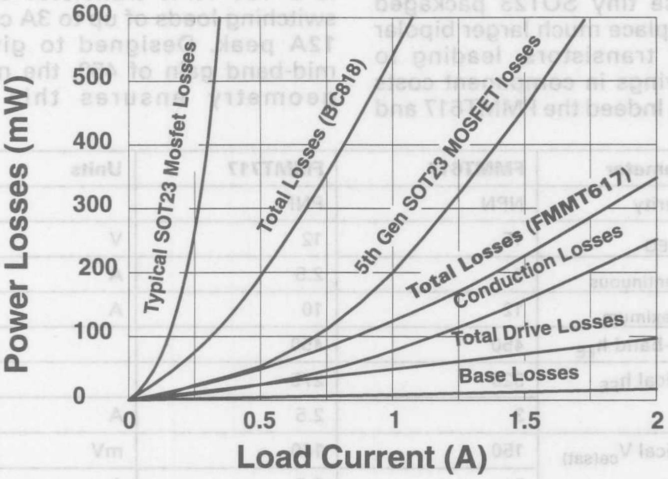


Figure 2
Graph of Power Losses (mW) vs Load Current (A).

However, the necessary base current must be taken from the 4.8V supply, incurring additional base drive losses in the driver circuit, raising the driver losses to a level comparable with conduction losses. To even approach such low conduction losses, competitive bipolar types would require far more base drive than the FMMT617 and hence cause base drive losses many times higher. To demonstrate this, the total losses of a FMMT617 and a competitive driver transistor (BC818) are also charted in Figure 2.

Using MOSFETs would eliminate the drive power losses, but as can be seen by reviewing the performance of the best MOSFET presently available in SOT23 (see Figure 2), the on-resistance makes conduction losses very much higher than the total losses obtained using the FMMT617. Furthermore, battery supplies rarely provide sufficient gate-drive voltage for the MOSFETs to obtain their minimum on-resistance values, a factor that rapidly gets worse as the batteries reach their end of life voltage.

The low losses obtained using the FMMT617 in this motor driver circuit can only be matched by competitive bipolar or MOSFET devices in much larger packages, e.g. TO220, D-PAK, the best performance examples of SOT223 or SO8 etc.

These alternatives are considerably more expensive, eat up far more PCB area, are more difficult to mount and often unable to match the reliability of SOT23 packaged devices. The motor driver circuit demonstrated is an example of the kind of high current, efficiency sensitive applications for which the FMMT617 and FMMT717 are particularly advantageous. Following are further application examples where these devices are ideally suited.

Motor Drivers

Providing bi-directional motor drivers for battery powered equipment requires either half-bridge controllers with centre-tapped battery packs or a full 'H'-bridge circuits with untapped batteries. Half-bridge controllers are simple and can be very cost-effective, but for a given motor power their drivers must pass twice the current of equivalent 'H'-bridge circuits. Also, as the operating voltage of the motor is only half of the total supply, the saturation voltage of the driver transistors must be kept to a minimum to maintain efficiency and battery life. The high current capability and low saturation voltage of the FMMT617 and FMMT717 make these transistors ideal for half-bridge controllers.

The circuit shown in Figure 3 is suitable for motors with peak currents up to 2.5A, giving saturation losses of only 90mV at 1A and 250mV at 2.5A. Using only 25mA base drive for the PNP and 15mA for the NPN, (current levels easily supplied by servo controllers and many logic ICs without the need for buffer stages), base drive losses are even smaller than conduction losses. Total conduction and driver losses are only a fraction of those obtained using industry standard transistors such as the BC818 and BC808, ensuring most of the energy taken from the battery is supplied to the load. High reverse gain eliminates the need for catch diodes in this circuit to protect the drivers from their inductive load, thus further reducing component counts. This half-bridge motor driver utilising the FMMT617 and FMMT717 gives useful savings in component count and costs without compromising efficiency or battery life. Typical applications are in positional systems, linear motor drives, servo and actuator drivers and toys.

In applications where the motor is driven predominately in one direction, it is possible for half-bridge circuits to discharge it's two battery sections unequally, wasting capacity and money when only partly discharged batteries are discarded. For these situations, a full bridge circuit is preferable. Although full bridge circuits halve the motor current required for a given power, two driver transistors are in the motor current path so low saturation voltage is still very important. Consequently, the FMMT617 and FMMT717 are equally applicable to this circuit topology too. Figure 4 shows a full bridge circuit using the FMMT617 and FMMT717 that will drive a 2.5A peak motor bi-directionally with exceptionally low losses. Note again that the buffers and catch diodes required by some competitive solutions are not required with this driver circuit, enhancing it's cost-effectiveness.

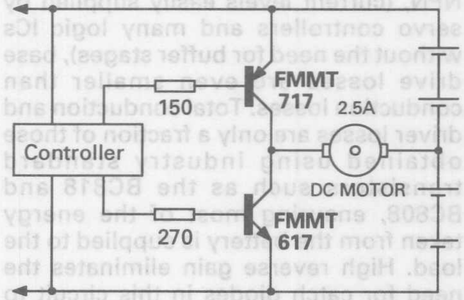


Figure 3
Half-bridge Motor Driver.

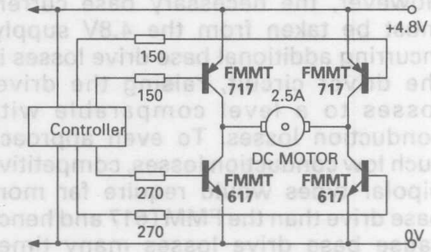


Figure 4
'H' bridge Motor Driver.

Active Power Switches (Including Mobile Communications and PCMCIA)

To maximise the endurance of complex battery powered products such as Mobile telecoms, Pagers, Laptop and Notebook computers etc., it is frequently necessary to switch the power supply of intermittently used circuit sections. To achieve low losses and ensure compliance with standard IC power supply specifications it is vital that these switches give very low voltage drops. For instance, 5% tolerance on a 3.3V supply equates to only 165mV. However, load currents can be high, for example the supply current demands of mobile phone RF output stages can peak at 2A, PCMCIA memory cards and hard disk drive modules can demand 300mA to 1A. To make the designer's life even harder, circuit constraints usually dictate that the +ve supply rail must be switched, making it difficult to use NPN bipolar transistors or N-channel MOSFETs which perform better than their P type counterparts. Meeting these stringent requirements usually necessitates the use of large and expensive P-channel MOSFETs or switch ICs.

The PCMCIA interface standard in particular adds to the cost and complexity as not only are it's supplies switched, they can also be set to several voltage levels, causing isolation problems that double the number of active switches needed.

Considerable savings in cost and component count can be made by substituting FMMT717 bipolar transistors for the alternative solutions currently used as supply switches. The low saturation voltage of the FMMT717 makes it an excellent low loss switch. Figures 5 and 6 show two PCMCIA interface supply switches. The PCMCIA interface standard provides two switched outputs, V_{cc} and V_{dd}. The supply V_{cc} must be switchable between 3.3V and 5V, and V_{dd} between 0V, 5V and 12V. The V_{cc} switch in Figure 5 is capable of sourcing 1A continuous and will supply peaks of over 3A. Controlled simply via logic drive signals, the typical voltage drop of either the 3.3V or 5V switch at 1A is around only 80mV. To achieve comparable performance using P-channel MOSFETs switch would require chips so large they would only fit in SO8, D-PAK, TO220 etc. and cost many times more than the SOT23 FMMT717 bipolar transistors. Also, when MOSFETs are used as the active switch element, two devices, one reversed and wired in series, must be used for the 3.3V switch to isolate the 3.3V input supply from the circuit's output when 5V out is selected. This is because the MOSFET's body diode conducts if the device becomes reverse biased, so a second device, (reverse wired) is required to block this effect. Not only does the use of MOSFETs for the 3.3V switch double its cost, it also doubles on-resistance and hence on-voltage drop. The FMMT717 can safely block a reverse voltage of 1.7V (5V-3.3V) hence no special measures are necessary.

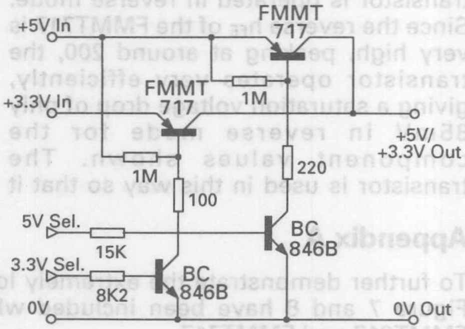


Figure 5
Typical Positive Line Switch - 1.

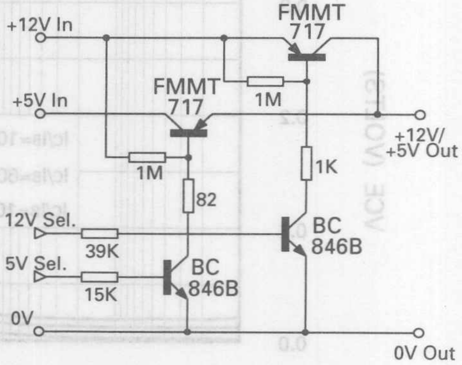


Figure 6
Typical positive line switch - 2.

The V_{dd} switch has been designed to source 1A from either the 5V or 12V outputs as with the Vcc switch but accepting slightly higher voltage drops. Although the Vdd circuit looks very similar to the Vcc switch, note that the collector-emitter leads of the 5V V_{dd} switch have been reversed. This transistor is operated in reverse mode. Since the reverse h_{FE} of the FMMT717 is very high, peaking at around 200, the transistor operates very efficiently, giving a saturation voltage drop of only 85mV in reverse mode for the component values shown. The transistor is used in this way so that it can isolate the 5V input line from the switches output when the 12V select line is activated. This requires the 5V switch to block a reverse voltage of 7V (12V-5V) and the FMMT717 must be connected as shown to guarantee that it can do this. P-channel MOSFETs could be used in this application but the cost and performance penalties cited against their use for the Vcc switch are equally applicable here. By exploiting the high reverse h_{FE} of the FMMT717, this PCMCIA V_{dd} switch gives a simple low cost circuit with excellent performance that is very hard to beat.

Appendix A

To further demonstrate the extremely low loss exhibited by the SuperSOT series, Figure 7 and 8 have been included which reproduce the $V_{CE(sat)}$ curves for the FMMT617 and FMMT717.

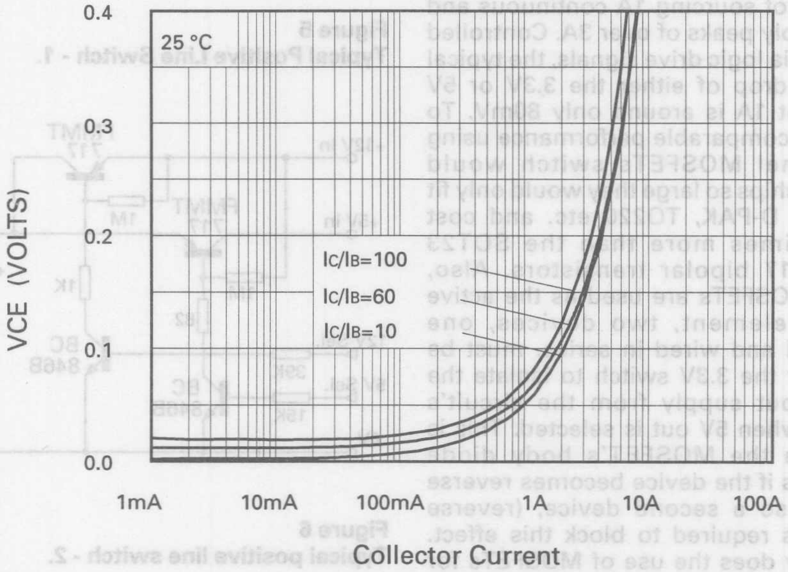


Figure 7
FMMT617 $V_{CE(sat)}$ v I_C.

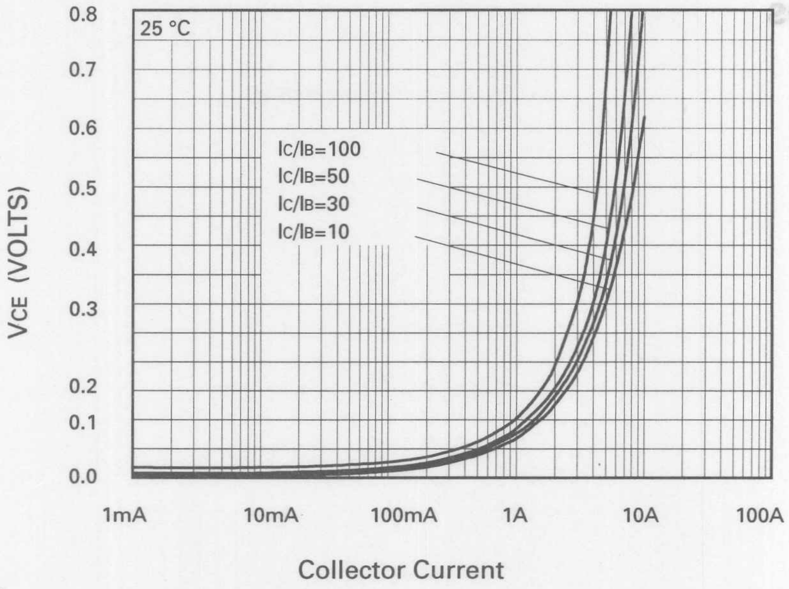


Figure 8
FMMT717 $V_{CE(sat)}$ v I_C

Notes

ZETEX

Application Note 16
Issue 1 March 1996

Automotive and Household Siren Driver Circuits

ZSD100 and Discrete 'H' Bridge Minimum Part Count Solution

David Brotton
Neil Chadderton

Many of today's modern alarm systems employ a moving coil loudspeaker as the siren element. To interface between the alarm system's processing unit and the speaker, the usual circuits consist of a signal generation IC and a discrete 'H'-bridge.

The signal generation circuit is often built around standard 555 timers, although other systems may use discrete transistor circuits, op-amp derived circuits or even small microcontrollers. The 555 configuration will consist of a slow running astable multivibrator, generating a ramp signal, which modulates a second multivibrator causing it to sweep over a range of frequencies.

The drive circuit usually employs TO126 or TO220 packaged power transistors, since the currents involved and the resulting power dissipation are both in excess of common smaller packaged products. The devices used normally require parallel collector-emitter diodes, to divert destructive reverse transients generated by the inductive load, and base emitter resistors to provide a path for any transistor leakage current. This factor is particularly important with usual TO126/TO220 products as their poor $V_{CE(sat)}$ performance can lead to a significant temperature rise. Some circuits use Darlington TO220's which can reduce the component count if the devices include integral collector-emitter diodes however, the Darlington configuration produces a high on-state voltage which causes excess dissipation. This type of circuit requires base emitter resistors and also reduces the power delivered to the load.

The result of the above is a complex circuit, with a high component count, requiring excessive board space and a consequential high cost. Figure 1 indicates a typical, and complex, solution to the circuits described above.

The Zetex total siren driver solution addresses all these problems in one by utilising a custom designed ASIC, which has a minimal pin count and uses few external components, and the superior performance of it's discrete transistors. The result is a low cost, minimum component count solution which still maintains the performance advantages expected from a Zetex product. Figure 2 shows the Zetex circuit solution to that shown in Figure 1.

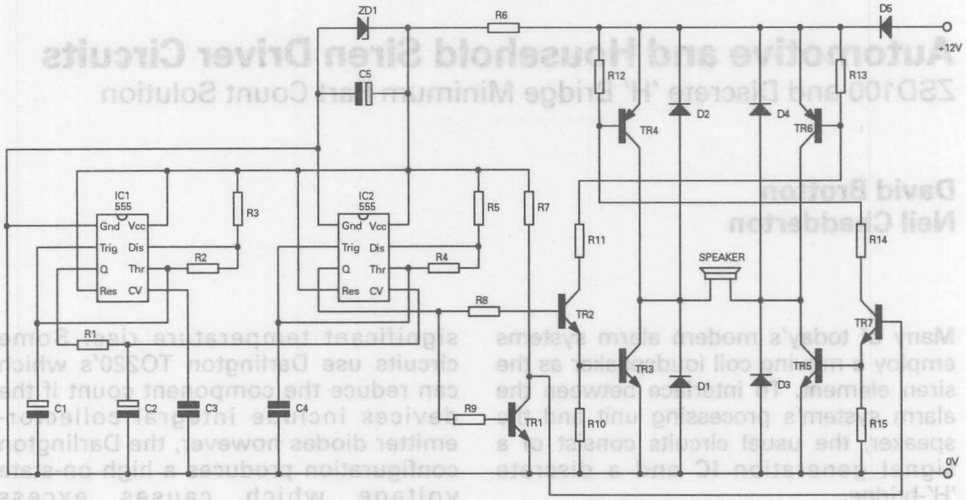


Figure 1.
Typical Circuit based on a dual 555 Configuration.

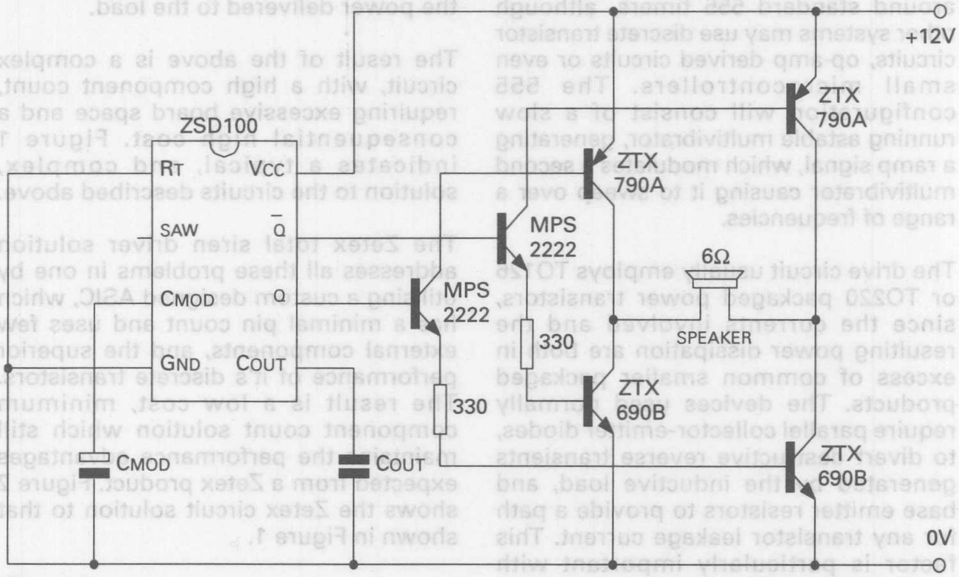


Figure 2
Zetex Equivalent Minimum Component Count Solution.

H-bridge circuits

Zetex recommend the use of their Super E-Line transistors in this application. These TO92 style components have three important features which are exploited in addressing the problems described above.

Zetex discrete transistors use an exceptionally efficient Matrix die geometry to achieve very low saturation voltages, $V_{CE(sat)}$, which is ideal for high current switching applications. This means for the typical operating currents required, normally in the region of 2A, there is very low dissipation and therefore minimal temperature rise. For example the ZTX690B devices shown in the diagram, with I_C=2A and I_B=20mA have a saturation voltage of only 300mV.This performance factor means that the need to divert leakage currents at high temperatures is removed and so the usual base emitter resistors are no longer required. This may be particularly important in automotive applications where the operating environment can produce very high temperatures.

The devices chosen for this application also possess an excellent reverse beta. Using the previous example, the ZTX690B has a typical peak beta of 125 in the reverse mode. This means that these transistors are still operating even with reverse transients such as those generated when driving an inductive load. The benefit of this is that for siren driving applications the need for collector-emitter protection diodes is removed.

A further benefit of the transistor matrix design and low saturation voltage is that the devices can be housed in a very small package. The Zetex E-Line package is smaller than a TO92 yet the performance of the devices matches that of similar TO126 and TO220 components. The result of this is equivalent or superior performance within a much smaller space.

The circuit shown in Figure 2 has been designed to operate with a 6 ohm load impedance or higher. If a lower load impedance is required, say 4 ohms, then changing the transistors to a selection from the ZTX850 and ZTX950 series will provide up to 5 amps drive current. The circuit we have shown has been tested and is 98.5% efficient- the resultant siren produces more than an ear piercing 120dB output.

The drive circuit for this application has thus been minimised in terms of complexity, takes up less space, performs better and offers significant cost reductions.

ZSD100 Siren Driver ASIC

The ZSD100 is a linear bipolar ASIC designed specifically to complement the drive circuits described above. The device utilises an 8 pin dual in line package and a minimum number of external components to generate a frequency swept signal suitable for either static or automotive security alarm systems. Figure 3 shows a schematic of the device.

As with other circuit approaches the device consists of a low frequency

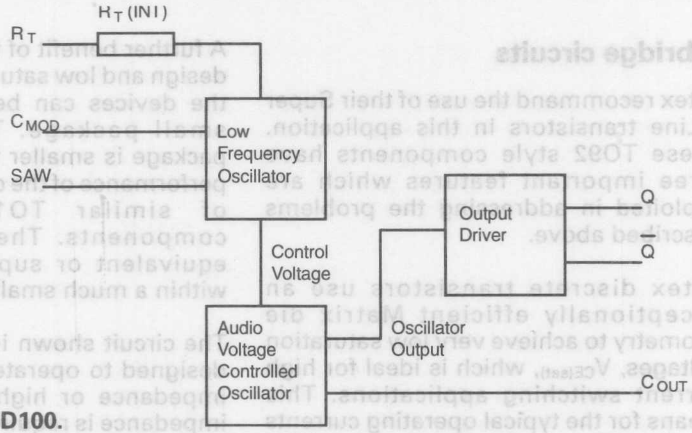


Figure 3
Schematic of the ZSD100.

oscillator, which generates a control voltage to modulate an on-chip voltage controlled oscillator.

The output stage of the device consists of a complementary driver which works equally well whether driving MOSFET's or IGBT's as alternatives to the bipolar transistor circuits shown. The output configuration enables not only the drive requirements required for full bridge circuits, as described above, but also single ended drive circuits such as would be required for piezo transducers, the other popular alarm sound generating device. This circuit is shown in Figure 4.

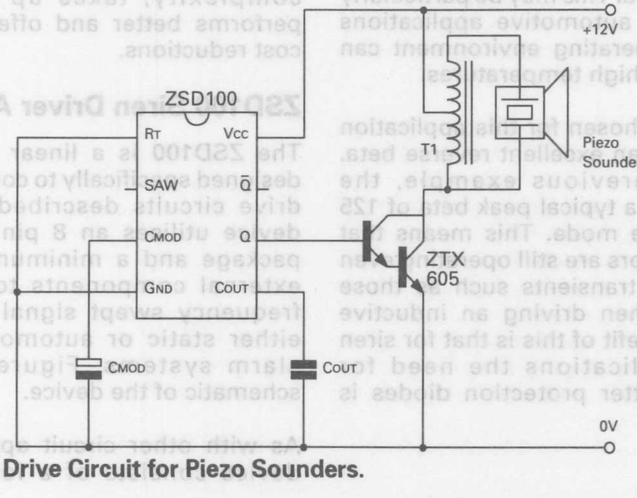


Figure 4
Single Ended Drive Circuit for Piezo Sounders.

The frequency programming of the device can be achieved with only two capacitors, although an extra pin is provided which provides improved control of both the low and high frequency oscillators if an external resistor is used, see Figure 5. Selection of the appropriate capacitors gives a low frequency sweep in the range 0.1 to 10Hz and an output frequency modulated between 100Hz and 10kHz.

The external resistor pin can also perform another function in that if it is driven from a logic source, which could be the alarm microprocessor, it can act as a disable by inhibiting all the driver functions. (Figure 6.)

One further feature of the ASIC is that the modulation waveform can be simply selected as either ramp or sawtooth, by using the SAW pin, giving variety in the properties of the siren output waveform. With battery operation in mind the IC can operate from 4 up to 18 volts consuming only 1µA in sleep mode. Another power saving property of the IC is the built in anti cross-conduction delay which, in full bridge circuits, prevents both arms of the bridge conducting simultaneously.

Surface mount packaging

The circuits illustrated make use of through hole products. All the components mentioned however are available as surface mount alternatives making for even greater space saving. E-Line is complemented by SOT23 and SOT223 products whilst the IC can be made available in an SO8 package as well as the DIL8.

Summary

The circuit described is only one example of the possible configurations of alarm circuit solutions produced from Zetex discrete and IC technologies. For whichever configuration is required the Zetex solution will offer reduced component count, space saving and cost saving whilst maintaining superior performance

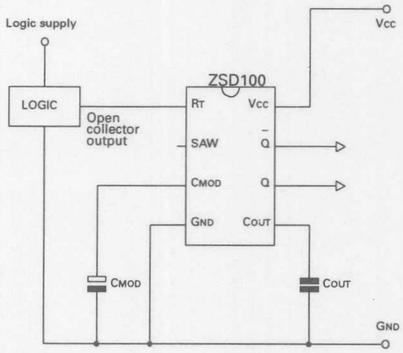


Figure 5
Improved Frequency Control.

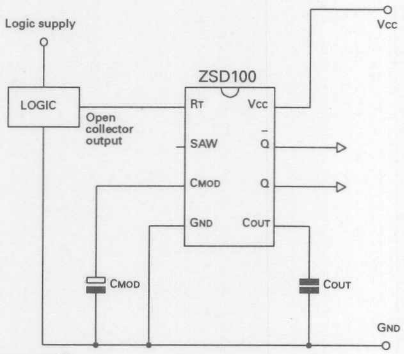


Figure 6
Logic Controlled Inhibit.

Notes

Application Note 17
Issue 2 March 1996

Emergency Lighting Systems and Battery Powered Fluorescent Lighting

High Current TO92 Switching Transistors Provide 87% DC-AC Conversion Efficiency

David Bradbury
Neil Chadderton

Introduction

Emergency lighting systems are frequently employed as a required safety feature within business premises. These systems comprise essentially of a control circuit, a battery pack, a trickle charging circuit, and either an in-built incandescent lamp, or an inverter that will allow the battery supply to drive the existing fluorescent tubes. It is the latter version, and the requisite DC-AC inverter that are addressed within this application note.

Figure 1 presents a block diagram of a typical emergency lighting system. The control circuit monitors the mains supply, and if all is well, allows the charger to trickle charge the battery pack. In the event of the mains voltage failing, the controller then enables the inverter which provides sufficient power to a fluorescent tube to provide illumination.

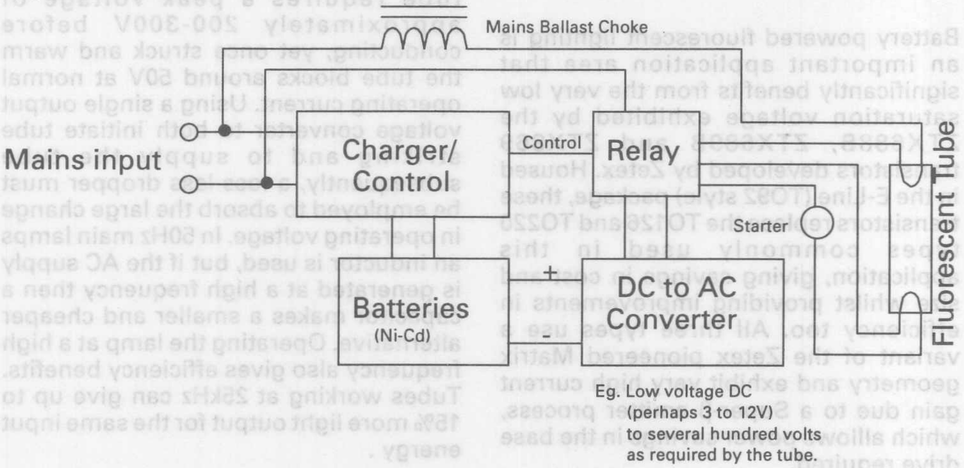


Figure 1
Typical Emergency Lighting System Schematic.

It should be noted that this tube is quite often a standard ceiling tube that would be otherwise powered by the mains supply.

Closely allied to this application, and indeed where similar circuits are employed, are illumination signs for entrance, exits and walkways.

Fluorescent tubes are employed in these applications because they are many times more efficient at converting electrical energy into light than conventional incandescent bulbs. This efficiency translates directly into extended battery life.

The number of series connected cells used in an emergency lighting system is kept to a minimum to optimise storage efficiency, reliability and cost. The resultant low output voltage of the battery pack does however make it more difficult to design high efficiency converters.

Battery powered fluorescent lighting is an important application area that significantly benefits from the very low saturation voltage exhibited by the ZTX688B, ZTX689B and ZTX869 transistors developed by Zetex. Housed in the E-Line (TO92 style) package, these transistors replace the TO126 and TO220 types commonly used in this application, giving savings in cost and size whilst providing improvements in efficiency too. All three types use a variant of the Zetex pioneered Matrix geometry and exhibit very high current gain due to a Super-β emitter process, which alllows power savings in the base drive required.

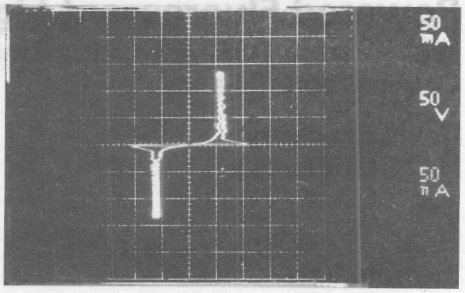


Figure 2
Current vs Voltage Characteristics for 8W Linear Fluorescent Tube.

Fluorescent Tube Characteristics

Figure 2 shows the V-I characteristics of an 8W fluorescent tube before and after striking. (An 8W tube being a common unit in illuminated door signs for example). Note that during striking the tube requires a peak voltage of approximately 200-300V before conducting, yet once struck and warm the tube blocks around 50V at normal operating current. Using a single output voltage converter to both initiate tube striking and to supply the tube subsequently, a loss-less dropper must be employed to absorb the large change in operating voltage. In 50Hz main lamps an inductor is used, but if the AC supply is generated at a high frequency then a capacitor makes a smaller and cheaper alternative. Operating the lamp at a high frequency also gives efficiency benefits. Tubes working at 25kHz can give up to 15% more light output for the same input energy .

The striking voltage of the tube increases both with age and decreasing temperature - so the open circuit output voltage of the converter needs to be set to around 500V (or higher, dependent on tube dimensions and characteristics) for reliable operation. The ballast capacitor value must be set to source about 120mA to supply the full tube power. In many emergency lighting systems, the output power of the inverter (and therefore the light output of the tube) is limited so to provide a longer operating time from the available energy source.

The circuits presented in this note however have been designed to operate at an output power of 8W. This serves to demonstrate the capability of the transistors and the basic topology, and can be used as a starting point for development of lighting inverters for a customer's specific application.

High Efficiency DC-AC Converters

The high voltage AC supply required by the fluorescent tube is generated using a push pull switching converter forced to run in synchronised mode by the inclusion of a supply inductor. This converter topology is also known as a resonant version of the popular push-pull power inverter developed by George Royer, and so is often termed a "Royer" Converter.

The following describes two fluorescent lamp circuits, both for 8W output, but designed for different input voltage ranges. Both circuits enable very high conversion efficiency. Each is designed to operate from just two series connected cells, either lead acid types providing 4V nominal or Ni-Cd/Ni-MH types at 2.4V. The same basic circuit is used for both voltage variants, but the component and transformer details are adjusted accordingly.

The circuit shown in Figure 3 can be used for the 2.4V or 4V supplies by selecting the appropriate component values from Table 1, or modified for any other supply voltage by consideration of the required transformer turns ratio.

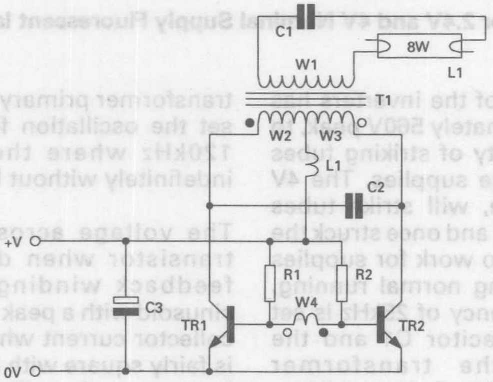


Figure 3
Resonant Push-pull Inverter for Low DC to High AC Voltage Inversion.

+V	2.4 Volts	4 Volts
R1	120Ω 0.5W	120Ω 0.5W
R2	120Ω 0.5W	120Ω 0.5W
TR1	ZTX869	ZTX689B or ZTX869
TR2	ZTX869	ZTX689B or ZTX869
C1	2.2nF 1000V Polypropylene	2.2nF 1000V Polypropylene
C2	0.47μF 100V Polyester	0.15μF 100V Polyester
C3	100μF 6.3V Electrolytic	100μF 6.3V Electrolytic
L1	25μH (25T, 1mm copper wire on 9mm dia 25mm long ferrite rod.)	60μH (35T, 0.71mm copper wire on 9mm dia 25mm long ferrite rod.)
T1	FX3440 cores with 0.55 spacer or FX 3670 cores with 0.65mm spacer. DT2484 coil former	FX3440 cores with 0.34 spacer or FX 3670 cores with 0.45mm spacer. DT2484 coil former
W1	500T 0.18mm neatly wound (first winding)	400T 0.18mm neatly wound (first winding)
W2 & W3	3T each, 0.5mm Bifilar wound (second and third windings)	4T each, 0.5mm Bifilar wound (second and third windings)
W4	3T 0.31mm (fourth winding)	3T 0.31mm (fourth winding)

NOTE: Use insulating tape between W1 and other windings. Core spacer must be made from a non-conducting material.

Table1
Component values for 2.4V and 4V Nominal Supply Fluorescent lamp Inverters.

The output voltage of the inverters has been set to approximately 560V peak, to provide the capability of striking tubes for low input voltage supplies. The 4V design for example, will strike tubes down to a 2.2V input, and once struck the lamp will continue to work for supplies down to 1.5V. During normal running, the operating frequency of 25kHz is set by the output capacitor C1 and the inductance of the transformer secondary winding W1. Before striking, or in the event of the tube being broken or removed, capacitor C2 and the transformer primary windings W2 & W3 set the oscillation frequency to a safe 120kHz where the circuit can run indefinitely without harm.

The voltage across each switching transistor when driven off (by the feedback winding) is a half wave sinusoid with a peak value of $\pi \times V_s$. The collector current when the device is on is fairly square with a ripple at twice the frequency of oscillation, controlled by the value of the series inductor L1. Figures 4 and 5 show these waveforms.

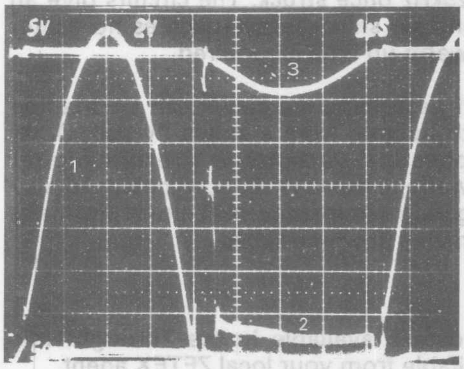


Figure 4
Inverter Operating waveforms for No Load
1. V_{CE}, 2. I_E, and 3. V_{BE},
2V/div, 0.5A/div and 5V/div respectively.
Fop ≈ 120kHz.

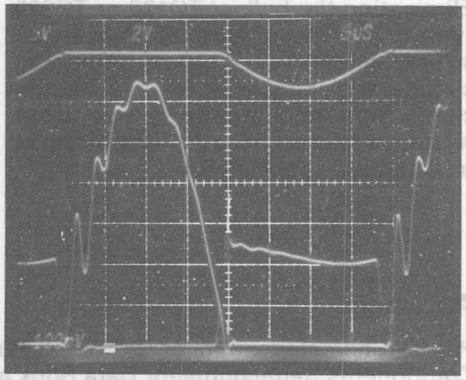


Figure 5
Inverter Operating waveforms for 8W Fluorescent Tube Load
1. V_{CE}, 2. I_E, and 3. V_{BE}.
2V/div, 1A/div and 5V/div respectively.
F_{op} ≈ 30kHz.

As the supply voltage is so small, low saturation voltage is critical in achieving good efficiency from the converter.

The designs have been optimised to meet the following key factors:-

1. Cost The designs operate from just two series connected Ni-Cd/Ni-MH or lead-acid cells - the less cells that are used in a battery pack, the cheaper and more volume efficient it will be. Also, replacing the TO220 type transistors normally used with E-Line ZTX689B or ZTX869 transistors reduces component and board size.

2. Battery Life The ZTX689B and ZTX869 transistors give by far the lowest saturation voltage of devices in their class. This translates directly to improved circuit efficiency and extended battery life. With most of the remaining losses occurring in the wound components, the efficiency of the 4V design is around 87% and the 2.4V design a very creditable 82%.

3. Supply Operating Range The 4V design will work for battery voltages in the range of 1.5V up to 8V. The 2.4V design from 0.95V to 6V. These wide operating ranges means that the circuits will withstand the high supply voltage that can occur with rapid charging, yet are capable of wringing the last ounce of charge from failing battery packs.

4. Reliability The designs give an enhanced reliability in several areas. The low power losses of the ZTX689B & ZTX869 transistors minimise temperature rises in the converter - important in reliability terms. Eliminating the bulk of TO126 or TO220 type transistors removes potential susceptibility to vibration. Also the circuits will withstand reverse battery connection and indefinite operation without a fluorescent tube - important in un-attended applications.

The designs operate at a frequency of around 100kHz during striking, falling to 25kHz once struck. The circuits give an instant start characteristic as no heater warm up time is required.

If necessary, the circuit can be adapted to operate from a single 2V cell. Other possible variants will drive higher wattage tubes at reduced power levels - giving emergency back-up of normally mains powered tubes.

Appendix

Partial Characterisation of ZTX688B. Full details available in the Through Hole Components Semiconductor Data Book available from your local ZETEX agent.

PARAMETER	SYMBOL	MIN.	TYP.	MAX.	UNIT	CONDITIONS.
Collector-Base Breakdown Voltage	$V_{(BR)CBO}$	12			V	I_C=100μA
Collector-Emitter Breakdown Voltage	$V_{(BR)CEO}$	12			V	I_C=10mA*
Emitter-Base Breakdown Voltage	$V_{(BR)EBO}$	5			V	I_E=100μA
Collector Cut-Off Current	I_{CBO}			0.1	μA	V_{CB}=10V
Emitter Cut-Off Current	I_{EBO}			0.1	μA	V_{EB}=4V
Collector-Emitter Saturation Voltage	$V_{CE(sat)}$			0.04 0.06 0.18 0.35	V V V V	I_C=0.1A, I_B=1mA I_C=0.1A, I_B=0.5mA* I_C=1A, I_B=50mA* I_C=3A, I_B=20mA*
Base-Emitter Saturation Voltage	$V_{BE(sat)}$			1.1	V	I_C=3A, I_B=20mA*
Base-Emitter Turn-On Voltage	$V_{BE(on)}$			1	V	I_C=3A, V_{CE}=2V*
Static Forward Current Transfer Ratio	h_{FE}	500 400 100				I_C=0.1A, V_{CE}=2V* I_C=3A, V_{CE}=2V* I_C=10A, V_{CE}=2V*
Transition Frequency	f_T	150			MHz	I_C=50mA, V_{CE}=5V f=50MHz
Input Capacitance	C_{ibo}		200		pF	V_{EB}=0.5V, f=1MHz
Output Capacitance	C_{obo}		40		pF	V_{CB}=10V, f=1MHz

*Measured under pulsed conditions. Pulse width=300μs. Duty cycle ≤2%

Power MOSFET Gate Driver Circuits using High Current Super-β Transistors

6A Pulse Rated SOT23 Transistors for High Frequency MOSFET Interfacing

Neil Chadderton

The power MOSFET is commonly presented and regarded as a voltage driven device, and as such there is a natural expectation that it can be driven from any pulse source, irrespective of that source's energy, or current capability.

This assumption is partly justified, if the system in question only pulses or switches the MOSFET at a low frequency, or in pure DC circuits, where the transistor may only be used in a toggled state. However, for typical switching frequencies from several kHz upwards, attention must be paid to the gate drive requirements to ensure efficient and "saturated" switching of the MOSFET. This must be considered as the gate-source (g-s) circuit is, to a first approximation, essentially a CR network; comprising the g-s capacitance, and the resistance of the metallic/silicon interconnects. To this network must be added the effective resistance, or source impedance of the gate driver circuitry, and for true assessments, consideration of the drain-gate (d-g) capacitance and the Miller effect. Due to this network, the g-s voltage follows an exponential curve as the C elements charge, and so, either sufficient time must be given to allow this voltage to reach it's target value (thus limiting the operating frequency and increasing the time spent in the linear region thereby producing high switching losses), or the "R" element must be minimised.

As a guide, the input capacitance of power MOSFETs ranges from a few hundred picofarads to tens of nanofarads. This capacitance is increased by the effective amplification of the drain-gate capacitance by the voltage gain of the circuit (Miller effect), such that the apparent capacitive component of the CR network assumes a value of 2 to 5 times the value of the datasheet stated C_{iss}. As this amplification effect is so circuit/bias condition dependent, a useful tool has been developed that considers the amount of gate charge required to meet a certain condition. Figure 1 shows a chart illustrating the gate charge required to switch the ZVN4306A (a 220mΩ, 1A continuous TO92 part), and this parameter's dependence on the Miller effect as the drain voltage increases.

As the operating frequency of switched mode power supplies increases, due to the need for less weight and product volume demands smaller inductors and capacitors, the MOSFET's required gate

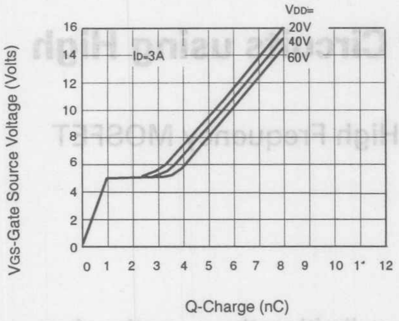


Figure1
Gate Charge Curves for the ZVN4306A; a 1A DC rated 220mΩ E-Line (TO92) Style MOSFET.

voltage must be driven to it's final value in as short a period as possible (within EMI constraints) to minimise switching losses. For a given value of required gate charge, this means that the current capability of the gate drive circuitry must be carefully considered.

As examples of the current required from the driver stage, and using the gate charge curves as a source:

i) A typical 100V, 300mΩ TO220 power MOSFET requires approximately 8nC, which for a switching time of say 20ns, leads to a current requirement of 400mA.

ii) A typical 500V, 900mΩ TO220 power MOSFET needs around 30nC, which could lead to a current requirement of 1.5A.

Obviously, paralleled MOSFETs are another concern, but a high current source could be used with the appropriate shared gate drive to reduce component count in this application also.

It is necessary therefore, that the gate driver circuitry acts as a low impedance voltage source, to enable the gate capacitance to be charged and discharged as quickly as possible. It must also have the capability of sourcing and sinking high transient gate currents - possibly several amps, in tens of nanoseconds. Standard logic family gates, and even the output stages of switch mode controller ICs are rarely able to provide this requirement and so would be unable to drive power MOSFETs in many applications. To provide an interface between the logic/PWM and the MOSFET, a high speed, high current capable (though not necessarily high power) buffer is therefore required.

The gate drive requirement is met by the complimentary emitter follower circuit shown in Figure 2, which should be

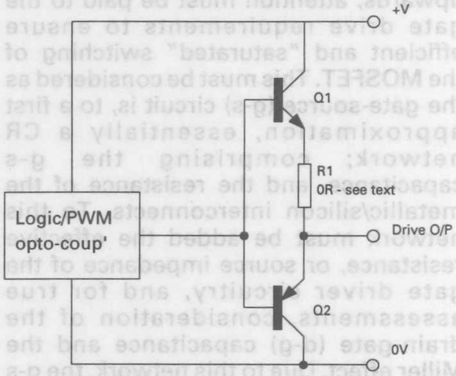


Figure 2
Complimentary Emitter Follower Gate Driver.

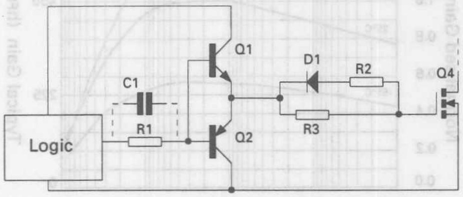


Figure 3
Complimentary Emitter Follower Gate Driver.

constructed with transistors possessing a high current capability (eg significant current gain at high collector currents), high F_T (as a benchmark to a fast switching capability), and ideally high gain. As the driver transistors only supply current while the capacitance is charging or discharging, the power capability (essentially determined by the package characteristics) is quite low, and can be tolerated by the smaller through-hole and surface mount packages.

By adjusting the amount of resistance in the charge path as shown, it is possible to delay the turn-on time. This may be necessary in some instances to prevent cross-conduction in push-pull output stages, or to decrease dV/dT to ensure compliance with EMI/RFI regulations.

This basic circuit can be adapted for different circuit topologies and performance requirements as shown in Figures 3, 4 and 5. Figure 3 shows another method of introducing unequal turn-on/turn-off times; Figure 4 shows a level shifted driver for a PMOS device; and Figure 5 a method of maintaining the correct drive level and drive phase, when deriving a control signal from a 5V logic based controller, by driving the emitter of a fast switching pre-driver transistor. This can either be a ZTX314 for a through-hole design, or a FMMT2369A for a surface mount version.

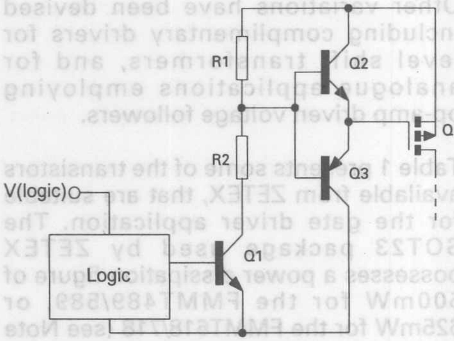
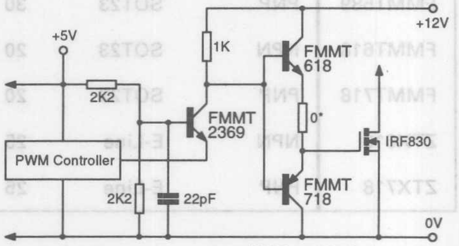


Figure 4
Level Shifted PMOS Gate Driver.



(* Set turn-on delay)

Figure 5
Complimentary Emitter Follower Gate Driver using Emitter Driven Switching Transistor to Retain Phasing.

Other variations have been devised including complimentary drivers for level shift transformers, and for analogue applications employing op-amp driven voltage followers.

Table 1 presents some of the transistors available from ZETEX, that are suitable for the gate driver application. The SOT23 package used by ZETEX possesses a power dissipation figure of 500mW for the FMMT489/589, or 625mW for the FMMT618/718 (see Note 1) - the latter being approximately twice that available from conventional SOT23 devices. This feature means that driver circuits previously effected with SOT89 packaged transistors can now take advantage of the smaller, lighter and more cost effective SOT23 range. The charts shown in Figures 6 and 7 illustrate the high current h_{FE} capability of the FMMT618 and FMMT718 transistors (also applicable to the through-hole versions - ZTX618 and ZTX718) which, with transition frequencies around 150MHz, guarantees high current drive integrity at the switching frequencies being demanded by modern power supply solutions.

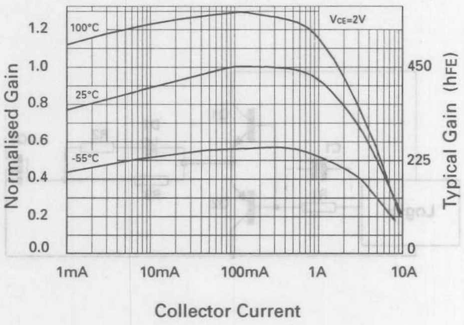


Figure 6
h_{FE} Profile for the FMMT618.

Device	Polarity	Package	B_{VCEO}	I_C (DC)	I_{CM}	$h_{FE(mb)}{}^2$
FMMT489	NPN	SOT23	30	1	4	200
FMMT589	PNP	SOT23	30	1	2	200
FMMT618	NPN	SOT23	20	2.5	6	450
FMMT718	PNP	SOT23	20	1.5	6	450
ZTX618	NPN	E-Line	25	3.5	10	450
ZTX718	PNP	E-Line	25	2.5	6	450

Note 1: When mounted on an industry standard 15 x15mm ceramic substrate. For an FR4 assembly, and a board of 1" x1", the FMMT618/718 series of "SuperSOT" transistors can achieve a Pd of 700mW.

Note2: h_{FE} (mb) - the value of the mid band current gain.

Table 1
Zetex Bipolar Transistors for MOSFET Gate Driver Applications.

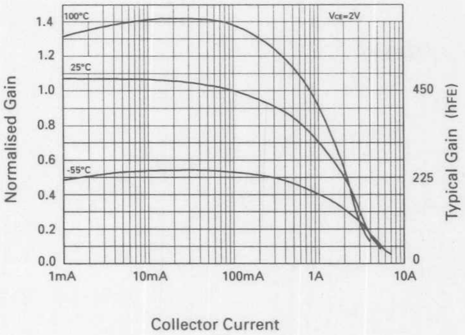


Figure 7
h_{FE} Profile for the FMMT718.

Figures 8 and 9 are oscillographs showing the response (albeit somewhat contrived for the purpose of illustration) of the g-s voltage and the resulting charge and discharge current provided by the complimentary emitter follower shown in figure 2. Figure 8 is for a 500V 3Ω part, while Figure 9 is for a 500V 900mΩ part. The lower trace in both cases being the control pulse to the driver stage, and the load is 2.4A resistive from a 350V supply. A peak pulse current of 400mA, and 1.6A being apparent respectively. A Tektronix current probe was used to measure the gate current in a 0.75" loop.

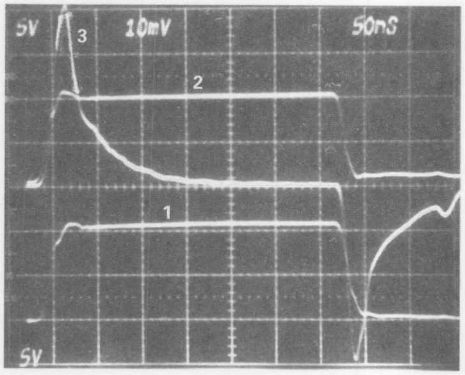


Figure 8
High Current Gate Driver with 500V 3Ω MOSFET; 2.4A Load.
1. Logic signal 5V/div., 2. G-S Voltage 5V/div., 3. G-S Current 100mA/div. Timebase at 50ns/div.

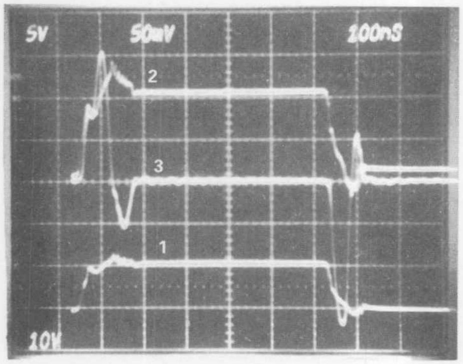


Figure 9
High Current Gate Driver with 500V 900mΩ MOSFET; 2.4A Load.
1. Logic Signal 10V/div., 2. G-S Voltage 5V/div., 3. G-S Current 500mA/div. Timebase at 100ns/div.

Notes

Application Note 19
Issue 2 November 1995

Component Reliability

Introduction

The reliability of Zetex components is rigorously monitored by the use of product qualification and on-going production testing. Testing is not done only to collect data, but this data is used to improve the reliability of future Zetex components.

In this note, the basic theory of reliability testing and failure rate calculations are covered, along with how Zetex carry out component qualification exercises and failure analysis, so as to ensure the on-going improvement of Zetex products.

Reliability Theory

Reliability can be defined as the probability of failure free performance of a required function, under a specified environment, for a given period of time. During this lifetime three distinct phases exist:

a) Infant Mortality:
Occurs in the first few hours of operating life. In this region, there is a high failure rate initially which then falls over the next few hours.

b) Random Failures:
Follows the infant mortality period and occurs over a long period of time. During this period there is a low, almost constant failure rate, with failures occurring due to random effects.

c) Wear out Region:
During this phase the number of failures begins to rise again as a result of devices reaching the end of their useful life.

A graphical representation of these three phases is shown in figure 1. This curve is frequently referred to as the "bath tub" curve.

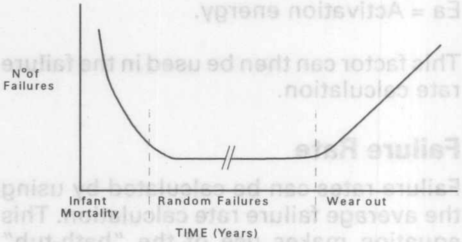


Figure 1.
Number of Failures with time.

Calculation of Failure Rate

Numerous concepts and mathematical models have been proposed to assess the reliability performance of semiconductor components. In this section, two parameters will be considered; thermal activation energy and failure rate.

Thermal Activation Energy

The thermal activation energy is used in determining the acceleration factor for a particular test when related to a lower standard operating temperature. The thermal activation energy can be determined for each failure mechanism and then substituted into the Arrhenius equation so as to calculate the thermal acceleration factor.

$$\text{Acceleration factor}=\exp[Ea/K(T_1^{-1}-T_2^{-1})]$$

where:

T_1 = Test temperature in degrees K.
T_2 = Standard operating temperature in degrees K.
K= Boltzmans constant.
Ea = Activation energy.

This factor can then be used in the failure rate calculation.

Failure Rate

Failure rates can be calculated by using the average failure rate calculation. This equation makes use of the "bath-tub" curve where the random region is assumed to be constant. Appropriate junction temperatures and thermal acceleration factors are used in the calculation along with a confidence limit (usually 60%) for the activation energy. The failure rate for a particular device is then calculated as:

$$\text{Failure Rate} = \frac{K(n) \times 10^9}{H}$$

where:
K(n) = Factor dependant upon total number of failures and the upper confidence limit chosen.
H = Equivalent number of device hours on test (i.e. utilising the acceleration factor calculated) to accumulate "n" failures.

The failure rate is thus quoted in FIT's with 1 FIT (Failure In Time) equating to 1 failure per billion hours of operation.

NOTE: FIT calculations for generic Zetex product (by package style) are included in appendix A.

Product Qualification

Qualification Tests

All new Zetex products and process changes are qualified using accelerated test methods, to ensure that the design or process change is reliable. These accelerated tests are used to model actual service conditions through increased temperatures, humidities, voltages etc. The following sections summarise these tests:

a) High Temperature Reverse Bias (HTRB)
The HTRB test is used to accelerate failure mechanisms which are thermally activated through the application of extreme temperatures and the use of

biased operating conditions. The temperature and voltage conditions used in the stress will vary with the product being tested.

b)Electrical Endurance and Power Cycling
These tests are performed to assess the quality of die attach and wire bond processes and is carried out with the product operating at maximum power, (with electrical endurance tests) and intermittent maximum power (for power cycling tests).

c) Humidity Tests
Three humidity tests are performed: (i) Temperature Humidity bias, (ii) Damp heat cyclic and (iii) Autoclave. In the temperature humidity bias test, an environment of 85°C, 85% relative humidity is established. The test is designed to measure the moisture resistance of plastic encapsulated devices. A reverse bias is applied to the device to create electrolytic cells which can then accelerate corrosion of the aluminium metalisation, if contamination is present.

The cyclic damp heat test is designed to examine the combined effects of temperature cycling, high humidity and temperature. This test is performed un-biased and uses an environment of 55°C and 95% relative humidity.

The autoclave test uses an environment of 121°C, 100% relative humidity and 15 psi. These conditions are used to test the devices' resistance to moisture penetration and the resultant effects of corrosion.

d) High Temperature Storage
This test is used to measure the stability of packaged devices to long term, un-biased storage. Typical failure mechanisms that will be accelerated by this test are die attach and wire bond related faults.

e) Temperature Cycling
Three temperature cycling tests are performed; (i) temperature cycling, (ii) thermal shock, (iii) resistance to solder heat.

With temperature cycling, the devices are cycled from -55°C to +150°C in an air environment with a total cycle from cold to hot taking one hour. This test is designed to accelerate the effects of thermal mis-match amongst the die/assembly components.

The objective of thermal shock testing are the same as that for temperature cycling. However, thermal shock presents additional stress conditions in that the device is exposed to a sudden temperature change created by the use of a liquid ambient.

The resistance to solder heat test is used to establish the devices' ability to withstand the temperatures seen during a board soldering process.

Qualification Programs

In order that products and process are designed in such a way as to ensure that reliable product is manufactured, new products and processes are thoroughly tested against a variety of the tests described in the previous section. Once released to the production departments, these products and processes are then continually assessed via the Zetex Reliability Monitor Program. This monitor program is designed to:

a) Provide a periodic evaluation of all products (see note below) on an on-going basis and,

b) To aid in the development of new products by establishing the current limitations.

NOTE 1: In order to ensure that a representative range of products are tested, the monitor program looks at all package styles on a monthly basis and product families on an annual basis, with the rate of testing commensurate with the volume of a particular product family being manufactured by the manufacturing group.

NOTE 2: Reliability data can be requested at any time from Zetex. MTBF and FIT values can be calculated for all package styles and for most product types. However, should only limited data be available for a specific product, generic calculations can be provided.

Failure Analysis

The primary purposes of failure analysis are:

1. To close the loop between the customer and Zetex for customer returned parts.

2. To determine the cause for device malfunction and parametric degradation.

3. To provide data for the continuous improvement in the reliability of Zetex products and processes. Material and failure analysis forms an important part of Zetex's commitment to product and process improvement. Results of all analyses are fed back into the development stage of new products, thus ensuring a continuing improvement in product reliability.

Failure analysis of all devices follows a systematic flow which leads to determining the root cause of failure. This analysis flow, along with some of the associated analytical steps is shown in figure 2.

Once a failure mechanism has been identified, the results are recorded and action taken to eliminate the cause in future by means of an improvement or corrective action team.

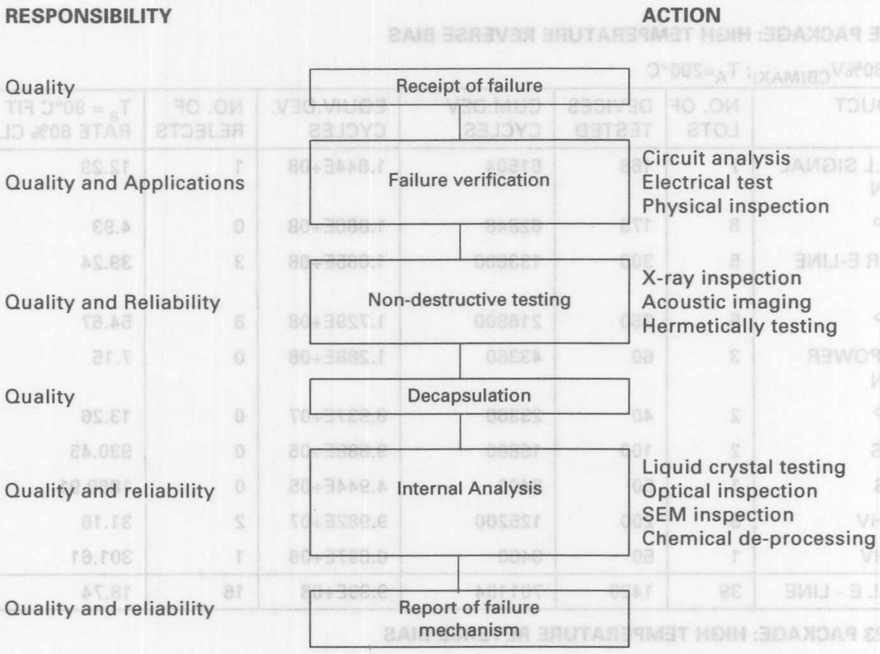


Figure 2
Failure Analysis Flowchart.

Appendix A

E-LINE PACKAGE: HIGH TEMPERATURE REVERSE BIAS

V_{CB}=80%$V_{CB(MAX)}$; T_A=200°C

PRODUCT	NO. OF LOTS	DEVICES TESTED	CUM.DEV CYCLES	EQUIV.DEV. CYCLES	NO. OF REJECTS	T_a = 90°C FIT RATE 60% CL
SMALL SIGNAL NPN	7	168	61504	1.644E+08	1	12.29
PNP	8	176	62848	1.866E+08	0	4.93
SUPER E-LINE NPN	5	300	133600	1.065E+08	3	39.24
PNP	5	250	216800	1.729E+08	8	54.67
MED POWER NPN	3	60	43360	1.288E+08	0	7.15
PNP	2	40	23360	6.937E+07	0	13.26
NMOS	2	100	16800	9.888E+05	0	930.45
PMOS	1	50	8400	4.944E+05	0	1860.91
NPN HV	3	200	125200	9.982E+07	2	31.16
PNP HV	1	50	8400	6.697E+06	1	301.61
TOTAL E - LINE	39	1420	701104	9.39E+08	16	18.74

SOT-23 PACKAGE: HIGH TEMPERATURE REVERSE BIAS

V_{CB}=80%$V_{CB(MAX)}$; T_A=150°C

PRODUCT	NO. OF LOTS	DEVICES TESTED	CUM.DEV CYCLES	EQUIV.DEV. CYCLES	NO. OF REJECTS	T_a = 90°C FIT RATE 60% CL
NPN SMALLSIGNAL	1	38	38000	8.853E+06	0	103.93
MED POWER	1	38	38000	8.853E+06	1	228.18
NPN SWITCH	2	115	50936	9.614E+06	2	323.49
NPN SUPERSOT	1	50	8400	4.944E+05	0	1860.91
PNP SUPERSOT	1	77	77000	4.597E+06	0	197.9
NMOS	7	484	324432	6.140E+06	2	162.87
PMOS	1	50	8400	4.944E+05	0	1860.91
NPN HV	5	238	71600	1.083E+07	6	678.67
PNP HV	5	341	193736	1.802E+07	1	112.11
ZENER DIODES	2	79	76000	1.771E+07	0	51.96
SWITCH DIODES	4	195	68536	1.339E+07	0	69.7
TOTAL SOT-23	30	1675	886440	1.078E+08	12	126.11

SOT-223 PACKAGE: HIGH TEMPERATURE REVERSE BIAS

V_{CB}=80% $V_{CB(MAX)}$; T_A=+150°C

PRODUCT	NO. OF LOTS	DEVICES TESTED	CUM.DEV CYCLES	EQUIV.DEV CYCLES	NO. OF REJECTS	T_a = 90°C FIT RATE 60% CL
NPN SWITCH	1	50	8400	4.944E+05	0	1860.91
NPN SUPER SOT223	3	150	116800	6.874E+06	1	293.85
PNP SUPER SOT223	5	300	183600	1.081E+07	1	186.94
NPN MED POWER	1	50	8400	4.944E+05	0	1860.91
NPN HV	2	100	108400	6.38E+06	0	144.2
NMOS	4	250	125200	7.369E+06	3	567.27
TOTAL SOT223	16	900	550800	3.242E+07	5	194.34

SOT-223 PACKAGE : ELECTRICAL ENDURANCE

T_A =+25°C, V_{CE} = 100% RATED V_{CE}: PD = MAX RATED POWER

PRODUCT	NO. OF LOTS	DEVICES TESTED	CUM.DEV CYCLES	EQUIV.DEV CYCLES	NO. OF REJECTS	T_a = 60°C FIT RATE 60% CL
NPN SWITCH	2	150	116800	9.695E+07	2	32.08
NPN SUPER SOT223	5	300	183600	1.145E+07	0	80.33
PNP SUPER SOT223	3	150	116800	9.695E+07	0	9.49
NPN MED POWER	2	100	58400	4.847E+07	0	18.98
PNP MED POWER	3	150	25200	1.572E+06	2	1978.44
NPN HV	1	50	8400	5.240E+05	4	10019.43
PNP HV	1	75	48376	1.014E+06	0	907.2
NMOS	1	50	50000	3.119E+06	0	294.97
TOTAL SOT-223	17	950	559200	4.641E+08	8	20.36

ELECTRICAL ENDURANCE : E-LINE PACKAGE

T_A = +25°C, V_{CE} = 100% RATED V_{CE}: PD = MAX RATED POWER

PRODUCT	NO. OF LOTS	DEVICES TESTED	CUM.DEV CYCLES	EQUIV.DEV. CYCLES	NO. OF REJECTS	T_a = 60°C FIT RATE 60% CL
SMALL SIGNAL NPN	27	2766	572016	9.155E+07	7	91.75
PNP	20	1605	305416	6.991E+06	6	1051.4
NPN SWITCH	2	118	91376	3.376E+05	0	2725.4
PNP SWITCH	4	225	145128	7.238E+06	0	127.1
SUPER E - LINE NPN	6	375	261928	2.607E+09	0	0.35
PNP	7	374	139376	1.284E+09	2	2.42
MED POWER NPN	13	964	275104	1.378E+09	3	3.03
PNP	5	340	148720	1.014E+09	1	1.99
NMOS	9	478	223408	1.111E+08	0	9.00
PMOS	2	175	106776	6.804E+07	0	13.52
NPN HV	4	290	48720	8.926E+07	0	10.31
SCHOTTKY DIODE	1	93	51400	9.152E+04	0	10052.01
TOTAL E - LINE	103	7899	2385496	6.663E+09	19	3.12

ELECTRICAL ENDURANCE: SOT-23 PACKAGE

T_A =+25°C, V_{CE} = 100% RATED V_{CE}: PD = MAX RATED POWER

PRODUCT	NO. OF LOTS	DEVICES TESTED	CUM. DEV CYCLES	EQUIV.DEV CYCLES	NO. OF REJECTS	T_a = 60°C FIT RATE 60% CL
SMALL SIGNAL NPN	9	540	355496	3.255E+07	0	28.27
PNP	9	575	275480	2.657E+07	0	34.63
NPN SWITCH	9	590	352144	2.310E+07	1	87.46
PNP SWITCH	4	222	100528	7.514E+06	3	556.31
SUPER SOT NPN	2	100	16800	1.316E+07	0	69.93
PNP	1	50	8400	6.578E+06	4	798.08
MED POWER NPN	6	365	174472	7.680E+07	0	11.98
PNP	8	531	322248	2.160E+08	0	4.26
NPN HV	2	113	54760	4.092E+06	0	224.78
PNP HV	4	201	69544	1.115E+07	0	82.52
ZENER DIODE	20	1353	1355160	1.094E+07	1	184.55
SWITCH DIODE	18	924	401504	3.723E+06	2	268.21
TOTAL SOT-23	93	5614	3494936	4.401E+08	11	28.63

PACKAGE INTEGRITY TEST RESULTS

HIGH TEMPERATURE/HIGH HUMIDITY

PACKAGE	NO. TESTED	CUMULATIVE DEVICE HOURS	FAIL	% FAIL
E-LINE	899	4.08E+05	2	0.22%
SOT-23	1749	6.43E+05	7	0.4%
SOT-223	1049	4.75E + 05	18	1.72%

TEMPERATURE CYCLING

PACKAGE	NO. TESTED	CUMULATIVE DEVICE CYCLES	FAIL	% FAIL
E-LINE	800	2.59E+05	4	0.5%
SOT-23	1151	5.42E+05	2	0.17%
SOT-223	950	5.59E+05	6	0.63%

THERMAL SHOCK

PACKAGE	NO. TESTED	CUMULATIVE DEVICE CYCLES	FAIL	%FAIL
E-LINE	1976	7.90E+05	0	0.0%
SOT-23	113	4.52E+04	0	0.0%

HOT STORAGE

PACKAGE	NO. TESTED	CUMULATIVE DEVICE CYCLES	FAIL	%FAIL
E-LINE	1681	1.68E+06	2	0.12%
SOT-23	1488	1.48E+05	0	0.0%

RESISTANCE TO SOLDER HEAT

PACKAGE	NO. TESTED	CUMULATIVE DEVICE CYCLES	FAIL	%FAIL
E-LINE	342	N/A	0	0.0%
SOT-23	890	N/A	0	0.0%

Application Note 20
Issue 1 April 1996

Magnetoresistive Sensors

Principles of Operation and Applications

Stefan Hübschmann
Matthias Schneider

This application note provides an overview of the Zetex range of Magnetoresistive sensors.

The layout of a typical magnetoresistive chip (parent device 174B) is shown in Figure 1, and is for example the chip used in the ZMY20 sensor. Thin film stripes are a characteristic feature of a magnetoresistive chip. These stripes are made by photolithography and consist of Permalloy ($Ni_{81}Fe_{19}$), a magnetic material evaporated on an oxidised silicon wafer. The electrical resistivity of the stripes is changed by a magnetic field H_y due to the magnetoresistive effect. The field H_y causes a rotation of the magnetisation in the stripe. This is shown in Figure 2. The resistivity R of a permalloy stripe depends on the angle between the directions of electric current (I) and magnetisation (M):

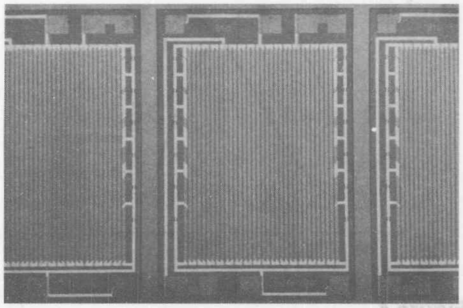


Figure 1
Magnetoresistive Magnetic Field Sensor, (Parent Device 174B).

$$R = R_0 + \Delta R_0 \cos^2\alpha$$

where ΔR_0 describes the strength of the magnetoresistive effect.

The maximum relative change of resistivity $\Delta R_0/R$ is approximately 2 to 3% for permalloy. The relationship between an external field H_y and angle α is

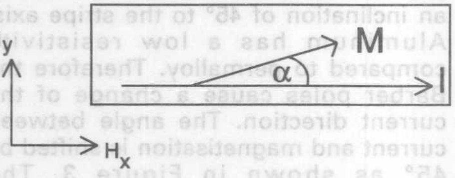


Figure 2
The magnetoresistive effect depends on the angle between the direction of electric current (I) and magnetisation (M). A rotation of the magnetisation in a permalloy stripe takes place when a magnetic field in the y-direction is applied. Without an external field the magnetisation is along the x-direction due the shape of the stripe.

determined by the geometrical dimensions of the stripe and the magnetic anisotropy of permalloy. This is taken into account by introducing a field H_0 that represents the demagnetising and anisotropic field. One obtains

$$\sin^2 \alpha = \frac{H_y^2}{H_0^2} \qquad \text{for } H \leqslant H_0$$

$$\sin^2 \alpha = 1 \qquad \text{for } H \geqslant H_0$$

The characteristic of a magnetoresistive stripe as a field sensor is:

$$R = R_0 + \Delta R_0 \left(1 - \frac{H_y{}^2}{H_0{}^2}\right) \qquad \text{for } H \leqslant H_0$$

A linear characteristic of the magnetoresistive sensor is required to measure a small magnetic field. The linear behaviour of the magnetoresistive sensor is achieved by using a "Barber pole" geometry. The stripes in Figure 1 are covered with aluminium bars having an inclination of 45° to the stripe axis. Aluminum has a low resistivity compared to permalloy. Therefore the Barber poles cause a change of the current direction. The angle between current and magnetisation is shifted by 45° as shown in Figure 3. The relationship between resistance and magnetic field is now

$$R = R_0 + \frac{\Delta R_0}{2} \pm \Delta R_0 \left(\frac{H_y}{H_0}\right) \sqrt{1 - \frac{H_y{}^2}{H_0{}^2}}$$

A linear characteristic of the sensor is given around $H_y^2/H_0^2 = 0$. The sign in this equation is determined by the inclination of the "Barber poles" (±45°) to the stripe axis. The characteristic of a sensor with and without Barber poles is presented in Figure 4.

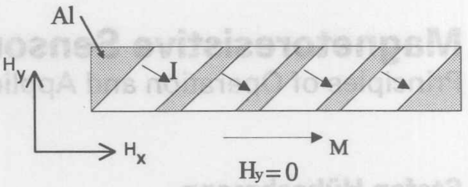


Figure 3
Covering the stripe with "Barber poles" consisting of aluminium changes the direction of the current. This does not influence the direction of magnetisation.

The stripes of the magnetoresistive chip are arranged as a meandering pattern. They form a Wheatstone bridge which is shown schematically in Figure 5. The applied voltage is V_b. Each half bridge consists of two resistors with different "Barber pole" orientations. The voltage between the resistors of a half bridge changes upon application of a magnetic field.

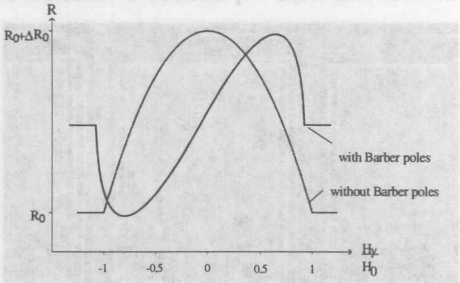


Figure 4
Characteristics of magnetoresistive sensors. The Barber pole structure enables a linear behaviour of the sensor for a small magnetic field.

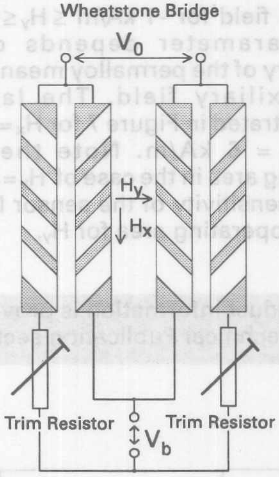


Figure 5
Wheatstone bridge of a magnetoresistive sensor with "Barber pole" structure. The bridge is balanced by laser trimming.

The resistance of one resistor increases, whilst the other resistor has a lower resistance due to the differing field characteristic. Adding a second half bridge with an opposite arrangement of "Barber poles" provides a Wheatstone bridge. The voltage difference V_0 is the output signal of the sensor. Each half bridge is trimmed to $V_b/2$ with an additional resistor in order to get an output voltage close to zero when no external field is applied. The trimming structures of the resistors in Figure 1 mark off the meander stripes on the left and right side of the chips.

Operating conditions and parameters

The shape of the stripe and the anisotropy of permalloy only define an axis along the x-direction for the magnetisation without external field H_y.

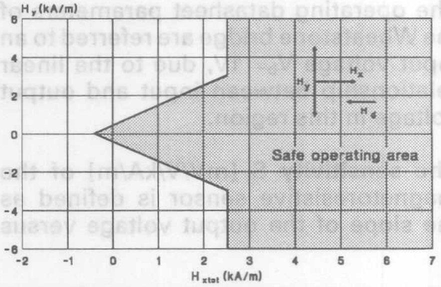


Figure 6
Safe operating area of ZMY20/ZMZ20.
$H_{xtot}=H_x + H_d$; T_{amb} = 25°C
H_d = disturbing field.

This means that in this state the stripe can have areas with a different direction of magnetisation (magnetic domains) and the sensor does not work in a stable way. A safe operation of the sensor is achieved by applying an auxiliary field H_x. This field defines the direction of the magnetisation. The range of H_y for safe sensor operation is determined by the strength of the auxiliary field. The safe operating area (SOA) of the sensor is demonstrated in Figure 6.
The field $H_{xtot} = H_y + H_d$ determines the allowed field values for H_y, where H_d is an external disturbing field in the x-direction.

There is no limitation for H_y in the case of $H_{xtot} \geq 2.6$ kA/m (ZMY20/ZMZ20). A small permanent magnet is sufficient to create the auxiliary field. The magnet can be glued on the sensor package (ZMZ 20/30 or ZMY 20/30). Another option is the ZMY20M which provides a very compact sensor including an integrated magnet, and is available in a surface mount package.

The operating datasheet parameters of the Wheatstone bridge are referred to an input voltage V_b= 1V, due to the linear relationship between input and output voltage in this region.

The **sensitivity** S [mV/V/kA/m] of the magnetoresistive sensor is defined as the slope of the output voltage versus external field for -1 kA/m $\leq H_y \leq$1 kA/m. This parameter depends on the geometry of the permalloy meander and the auxiliary field. The latter is demonstrated in Figure 7 for H_x = 3 kA/m and H_x = 6 kA/m. Note the small operating area in the case of H_y = 0 kA/m. A high sensitivity of the sensor leads to a small operating area for H_y.

Further details and complete Magnetoresistive Product information is provided in the Appropriate Technical Handbook: please see Technical Publication section.

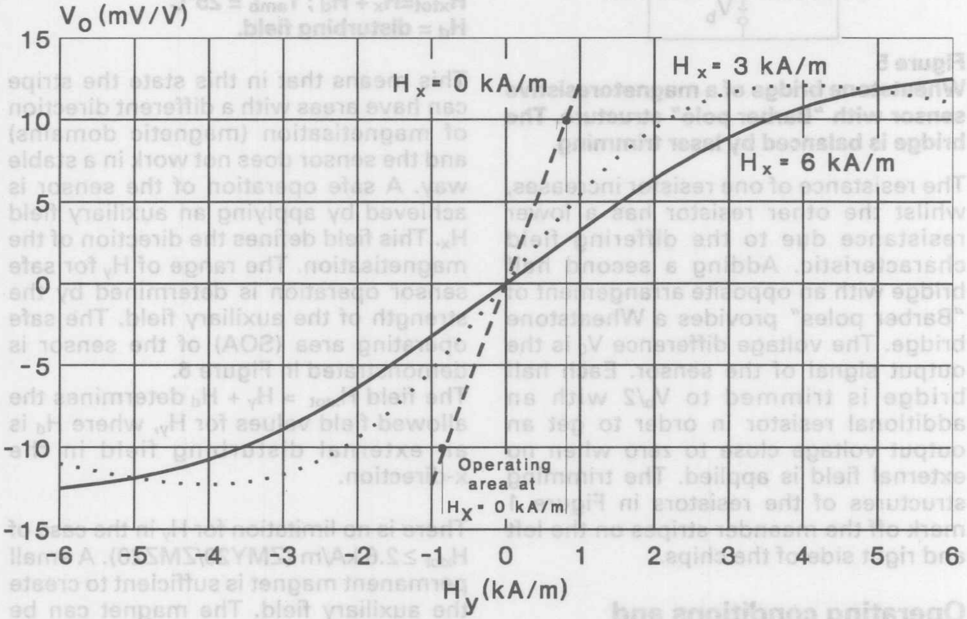


Figure 7
Sensor output characteristic of ZMY20/ZMZ20. The sensitivity of the sensor can be controlled by applying an auxiliary field H_X. This auxiliary field is necessary for sensor operation in a large field range. V_0=f(H_y); H_X-parameter;V_B=const;T_{amb}=25°C.

The Wheatstone bridge is balanced without the application of an external field ($H_y \leq 0.1$ kA/m). In this case the output voltage of the sensor is close to zero at room temperature. The deviation of the output voltage from zero is called the **offset voltage** V_{0ff}/V_b [mV/V]. The offset is caused by small geometric variations of the bridge which occur during the photolithographic process. The offset of the bridge is adjusted by laser trimming. The voltage output of each half bridge is $V_b/2$.

The **bridge resistance** R_{br} [%/K] of the magnetoresistive sensor depends linearly on temperature. The **temperature coefficient of bridge resistance** TCR_{br} [%/K] is positive. This is typical for metals. The **temperature coefficient of sensitivity** TCS [%/K] of the sensor is negative for V_B = const (TCS_V), because the strength of the magnetoresistive effect becomes smaller with increasing temperature. In the case of I_B = const (TCS_I), when the sensor is powered by a constant current supply, the temperature dependence of the sensitivity is reduced due to the linear relationship between input and output voltage. A higher bridge resistance caused by a rise in temperature leads to an increased applied voltage, partly compensating the change of sensitivity.

The Wheatstone bridge cannot fully compensate the temperature dependence of the resistors. The **temperature coefficient of offset voltage** TCV_{off} [µV/V/K] is due to local changes of resistivity in the permalloy thin film and photolithographic variations. This characteristic of the magnetoresistive sensor limits the measurement of small magnetic fields in a wide temperature range, especially in the case of static fields. Two sensors can be selected having a comparable temperature coefficient. The offset drift is partly eliminated by using the difference of the output voltages of both sensors. Another elegant way to avoid offset drift is to invert the direction of the auxiliary field and thus inverting the output voltage of the sensor. This can be done by small coils providing an auxiliary field that can change its direction.

The **hysteresis of output voltage** V_{0ffH}/V_b [µV/V] describes the accuracy of the magnetoresistive sensor. The magnetisation of the permalloy stripe is not completely homogenous. There are small areas of the meander, especially at the corners of the stripes, where the magnetisation is pinned and does not correctly follow the external field. The hysteresis is measured in a magnetic field loop, where H_y goes from -3 kA/m to 3 kA/m and back to 0 kA/m (H_x = 3 kA/m). V_{0ffH}/V_b denotes the shift of the offset voltage caused by this loop.

The **maximum range of output voltage** $\Delta V_0/V_b$ [mV/V] is defined as the difference of output voltage for $\alpha = 0°$ and $\alpha = 90°$, where α denotes the angle between current and magnetisation of the magnetoresistive stripe. This means that $\Delta V_0/V_b$ represents the strength of the magnetoresistive effect. This parameter decreases with temperature and determines the sensitivity of the sensor.

(An example of a typical Magnetoresistive sensor datasheet, is partially reproduced in Appendix B.)

Applications

[Please refer to appendix A which summarises some graphic examples of applications for magnetoresistive sensors].

Figure 8 shows a ZMC20 current sensor being used as a basis for an overcurrent trip switch used to protect power IGBTs within a motor driver system. The circuit reacts within 3µs to prevent latch-up related failure under transient/pulse conditions, and was built within a module measuring 35 x 20 x 25mm. An external 10kΩ preset is required for offset adjustment. Supply voltage is +5V ± 10% at 10mA; output is via an open collector transistor rated ar 1A, 20V; operating temperature range is 0 to 80°C.

Figure 9 provides a method for revolution measurement by reacting to a modulated magnetic field due to a rotating cog. The circuit gives a signal whose frequency is proportional to the rotational velocity of the cog, and a high level output for no rotation.

Figure 10 shows an application circuit for three-dimensional magnetic field observation. When the unit is enabled, it calibrates itself to the existing magnetic field of the earth, and then generates a warning signal if it is moved. The system employs three ZMY20 sensors (one for each dimension) and a CMOS EPROM microcontroller with an A/D converter. Similar circuits have been designed for automotive immobiliser/alarm systems that monitor the position of the vehicle by sensing the magnetic field of a movable permanent magnet. This magnet is necessary to shield the sensor

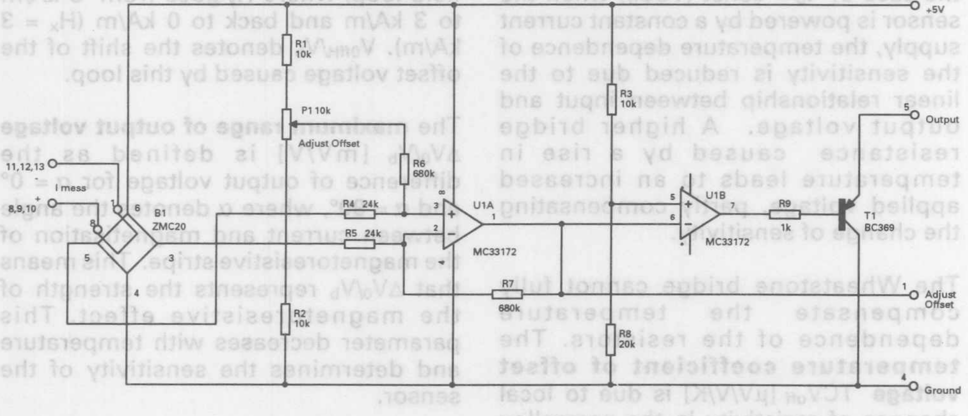


Figure 8
Overcurrent Switch using the ZMC20, for Protection of Power IGBTs.

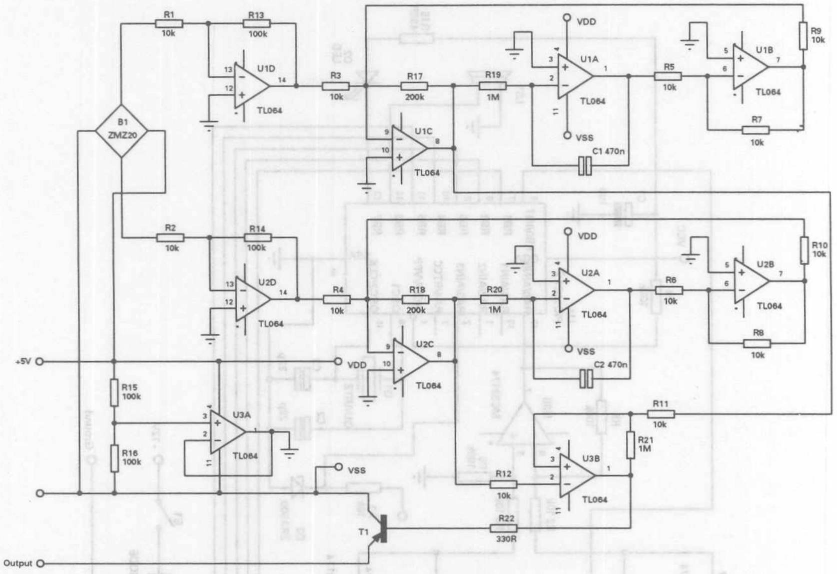


Figure 9
Sensor for Revolution Measurement.

from disturbing fields (generated by supply lines, car alternators etc.) Supporting software for these systems is available on request.

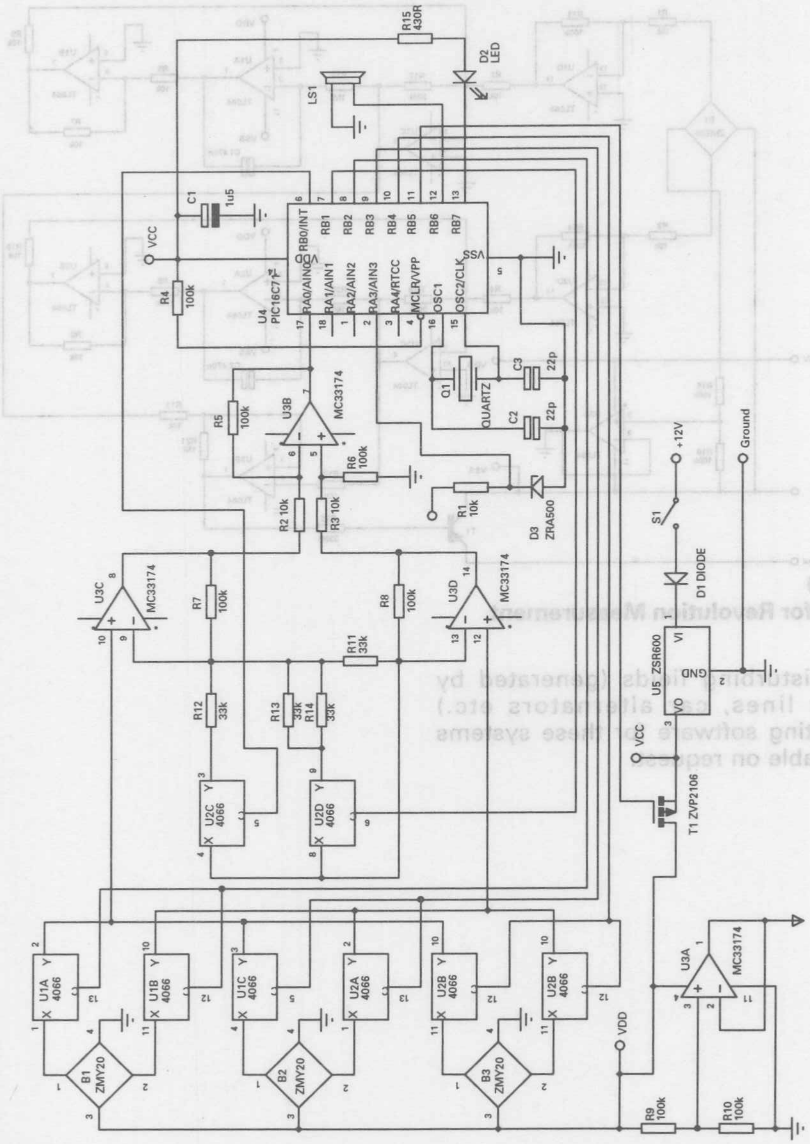
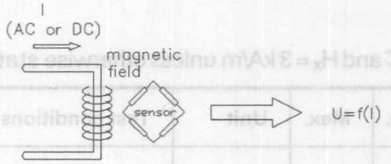
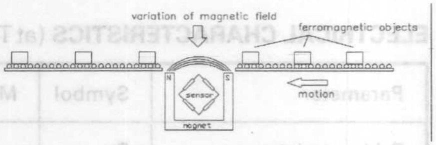


Figure 10
Sensor System used to Monitor Movement in the Earth's Magnetic Field.

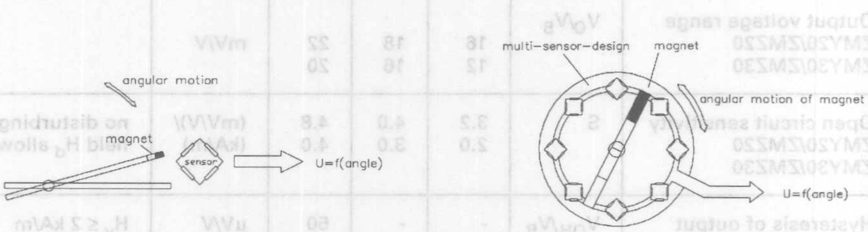
Appendix A
Magnetoresistive sensor Basic Function/Application Examples.



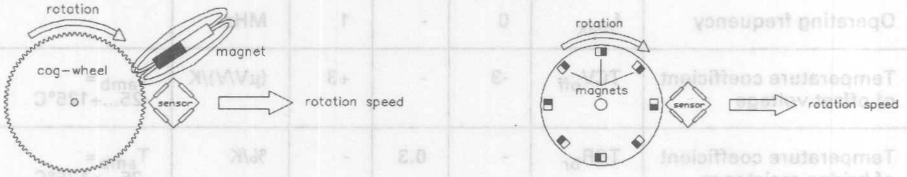
Measurement of Current (AC or DC)



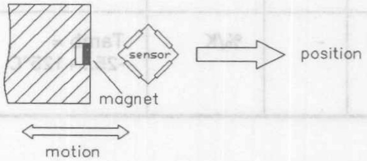
Detection of Ferromagnetic Objects



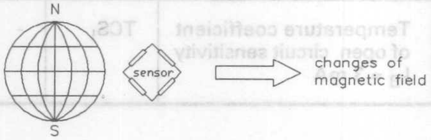
Measurement of Angular Position



Measurement of Rotation Speed



Position Sensor



Measurement of the Earth's Magnetic Field

Appendix B
Partial Characterisation for ZMY20/30, ZMZ20/30 Magnetoresistive Sensors.

ELECTRICAL CHARACTERISTICS (at T_{amb} = 25 °C and H_x = 3 kA/m unless otherwise stated)

Parameter	Symbol	Min.	Typ.	Max.	Unit	Test conditions
Bridge resistance ZMY20/ZMZ20 ZMY30/ZMZ30	R_{br}	1.2 2.0	1.7 3.0	2.2 4.0	kΩ	
Output voltage range ZMY20/ZMZ20 ZMY30/ZMZ30	V_O/V_B	16 12	18 16	22 20	mV/V	
Open circuit sensitivity ZMY20/ZMZ20 ZMY30/ZMZ30	S	3.2 2.0	4.0 3.0	4.8 4.0	(mV/V)/ (kA/m)	no disturbing field H_d allowed
Hysteresis of output voltage	V_{OH}/V_B	-	-	50	µV/V	$H_y \leq 2$ kA/m
Offset voltage	V_{off}/V_B	-1.0	-	+1.0	mV/V	
Operating frequency	f_{max}	0	-	1	MHz	
Temperature coefficient of offset voltage	TCV_{off}	-3	-	+3	(µV/V)/K	T_{amb} = -25...+125°C
Temperature coefficient of bridge resistance	TCR_{br}	-	0.3	-	%/K	T_{amb} = -25...+125°C
Temperature coefficient of open circuit sensitivity V_B = 5 V	TCS_V	-	-0.4	-	%/K	T_{amb} = -25...+125°C
Temperature coefficient of open circuit sensitivity I_B = 3 mA	TCS_I	-	-0.1	-	%/K	Tamb = -25...+125°C

ZETEX

Application Note 21
Issue 2 January 1996

Bipolar Transistor Considerations for Battery Powered Equipment

Leading to Efficiency and Competitive Advantages in Portable Systems

Neil Chadderton

Introduction

The last few years has witnessed an increasing trend towards portability, this no doubt being due to a waiting market, and the advances in the enabling technology within the digital domain. This in turn has produced impetus to the advancement of the analogue technologies, as the customer requirements dictate a move to integrated, lower cost, energy efficient products. These new technologies include higher capacity battery systems and a re-assessment of power management techniques. The new philosophy includes careful charge control to ensure maximum battery capacity and lifetime, and consideration of voltage and current ratings - leading to the design and characterisation of components specific to the application.

It is easy to assume that within say, a laptop computer or a mobile phone, that the circuit boards will be wholly populated with digital ICs, with little or no analogue circuitry. This is in fact far from the truth; as the control systems tend to incorporate more features, and microprocessors and microcontrollers move to higher speeds and lower operating voltages, the demands placed on the system power supplies, battery charge schemes, and circuit block power switches become more exacting. Portability implies smaller and lighter components, which is usually considered to be in opposition with higher power requirements. The analogue sections and consequently the switching devices used for these systems, must then be considered and chosen carefully to meet product objectives.

Zetex Semiconductors has developed a bipolar transistor technology that enables a range of devices ideally suited to many of the high current, efficiency conscious circuit concerns of todays battery powered and portable products. This technology, the Matrix Geometry, was initially introduced to effect small DC motor drivers for cameras, and has been greatly enhanced to provide a range of transistors un-matched by any other manufacturer. This range includes a 5A continuous device in the TO-92 style E-Line package; SOT223 rated up to 7A; the SuperSOT SOT23 series which includes a 3A continuous part; and the capability of producing Super-β transistors, thereby allowing cost-effective replacement of larger Darlington and MOSFET transistors.

This application note aims to provide a general overview of this Zetex bipolar transistor technology with particular respect to selection criteria, comparison against competing MOSFET solutions, and performance advantages for low voltage applications.

Bipolar Vs MOSFET

Bipolar technology has perhaps been somewhat overshadowed in recent years, particularly since the birth of the MOSFET. This is to be expected due to two main reasons. Firstly; each major new technological advancement brings a wealth of publicity, promotion and a vast exposure of new design methods and circuitry. Unfortunately, this same PR drive comes at some cost: it is by its nature very selective, and has led to a commonly held view that bipolar can always be replaced with a MOS based product, particularly when speed, cost effectiveness and on-state efficiency are of concern. This view is true **only some of the time.** If adopted in too general a fashion, this approach can lead to non-optimised products, with the usual market disadvantages in performance and cost. There is no single device, or single technology solution to every application. The second factor is a general stagnation of bipolar device research as semiconductor manufacturers move to the latest technologies. Zetex recognises that optimised bipolar products offer the best fit design option in many cases, and has pushed the technology to higher performance standards than any of it's competitors. This section shows that modern bipolar technology, can provide a credible alternative to MOS based designs, and in many cases is the preferred choice. This is not to decry the use of MOSFETs, but rather to demonstrate that each device technology has it's advantages **and** disadvantages, and that each new application should be judged individually, not on a wholesale basis.

The information shown in Table 1 provides a basic technology overview of important Bipolar and MOSFET characteristics. To do this, a comparison has been effected between a Zetex 3rd generation bipolar product such as the geometry/process used for the SuperSOT series, and a typical latest generation MOSFET device.

1. Bipolar still claims the highest **silicon utilisation** of any transistor technology. This is due to the pattern of current flow within the geometry. Optimised bipolar geometries force the majority of the current flow vertically through the structure. MOSFETs however, still need to channel the current initially in a lateral manner before conducting through the bulk of the device. This fact allows an optimised bipolar device to use a smaller area to exhibit a given level of on-state loss, or, put another way, a bipolar device can conduct higher levels of current for the same area of silicon. This smaller silicon area leads to smaller packaging options required to encapsulate that silicon (which is a main contributer to the final product cost) and of course smaller products. Figure 1 illustrates the Zetex pioneered Matrix geometry on which many of the leading edge products are based.

Another point worth considering is how the on-state loss varies with temperature. While the components of

Table 1
Bipolar and MOSFET Technology Overview.

Performance Feature	Zetex 3rd Generation Bipolar Transistor	Typical Latest Generation MOSFET
1. Silicon utilisation	Excellent	Moderate
2. Drive voltage	<1V (V_{BE})	<2.7V (V_{th}) to 5V
3. Drive power	Moderate (high β)	Very low
4. Speed	Fast	Very fast
5. ESD sensitivity	Rugged	Sensitive
6. Price	Moderate	Expensive

pure resistance will increase with increasing temperature, this may be compensated for (within a given drive condition) by decreasing threshold voltage (for the MOSFET) or increasing h_{FE} (for the bipolar). MOSFET datasheets typically show $R_{DS(on)}$ increasing by a factor of x1.7 to x2 over the operating temperature range. Bipolar devices, particularly low voltage variants as designed for battery powered applications, show a high degree of $V_{CE(sat)}$ temperature compensation, (please refer to Figure 2).

2. The amount of **drive voltage** required to activate, turn-on, enhance etc is an important characteristic of any semiconductor device. Different technologies either address the issue directly in terms of transconductance, (Gm, gFS), or by secondary effects, Eg. current gain.

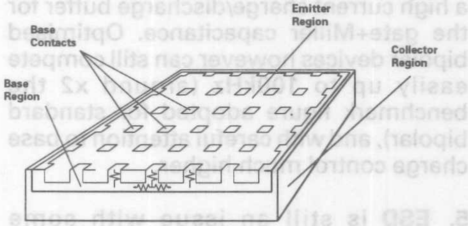


Figure 3
The Matrix Geometry
Distributed base resistance is minimised using a large matrix of base contact holes. By keeping the size of these holes small, little emitter area is lost and so active chip area is maximised.

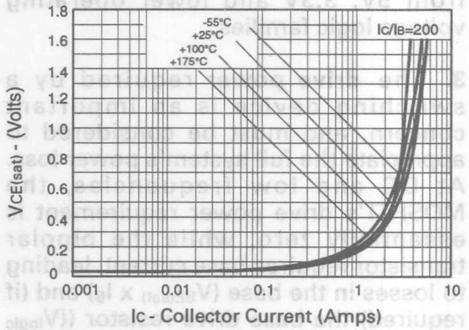


Figure 2
$V_{CE(sat)}$ vs Ic for ZTX788B, Illustrating Degree of Temperature Dependence.

Due to reducing system operating voltages, MOSFET vendors have reduced the threshold voltage, and therefore the gate-source voltage required for enhancement of the MOSFET channel. Even so, for full enhancement, to achieve datasheet and quoted resistance values, many standard MOSFETs can still require 10V or more, while low threshold devices need 4.5 to 5V. Many systems may not have this level of voltage drive available, so it is very important to fully assess the true level of on-resistance presented to the circuit, in order to understand the loss mechanisms.

For the bipolar device, the usual enabling parameter of concern is current gain, h_{FE} or β. However to assist this comparison exercise - in terms of drive voltage the bipolar transistor of course only requires a V_{BE} to promote collector-emitter conduction. For low currents this V_{BE} can be less than 400mV, perhaps rising to 1V or so at moderate to high currents, (please refer to Figure 3), thus allowing true logic level operation from 5V, 3.3V and lower operating voltage logic families.

3. The **drive power** required by a switching device is an important concern, and must be considered to appreciate the full system's power loss. At DC and low frequencies, the MOSFET's drive power requirement is essentially zero, while the bipolar transistor requires base current, leading to losses in the base ($V_{BE(sat)} \times I_B$) and (if required) the base drive resistor ($(V_{logic} - V_{BE})^2/R_B$). These losses can be minimised however, by employing devices with very high current gain.

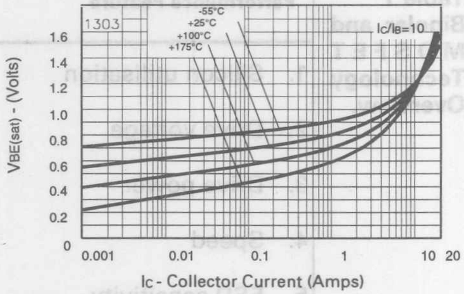


Figure 3
$V_{BE(sat)}$ or "Required Turn-on Voltage" for ZTX948.

These Super-β transistors possess typical mid-band h_{FE} values around 450 to 800 (dependent on device), which allows direct logic mA rated outputs to control single transistors which switch many amperes.

4. The **switching speed** capability is perhaps more straightforward - MOSFETs are operated routinely from 10s to 100s of kHz, and even to several MHz, although care must be taken with appropriate gate drive circuitry to ensure a high current charge/discharge buffer for the gate+Miller capacitance. Optimised bipolar devices however can still compete easily up to 100kHz (around x2 the benchmark figure adopted for standard bipolar), and with careful attention to base charge control much higher.

5. **ESD** is still an issue with some assembly contractors, though the compliance to safe handling standards, and investment in static safe environments will reduce this concern.

6. **Price**. As MOSFETs require more silicon area than bipolar parts for a given current capability, and MOSFET production processes demand state of the art alignment and etching, as well as more mask stages than bipolar, the difference in cost of manufacture and therefore the selling price, can be very significant.

7. The **reverse blocking** capability of a bipolar transistor depends on the state of the base terminal. If this is left open circuit/high impedance, as could be the case with a PNP in high side switch, then the B_{VEBO} parameter determines the reverse blocking voltage. MOSFETs cannot reverse block due to the inherent drain-source body diode. For some applications where reverse blocking is essential, MOSFETs can be configured as back to back pairs, such that each MOSFET blocks the body diode of it's partner. This does however double the on-resistance seen by the circuit. Figure 4 shows the typical B_{VECO} characteristic of a representative Super-β transistor.

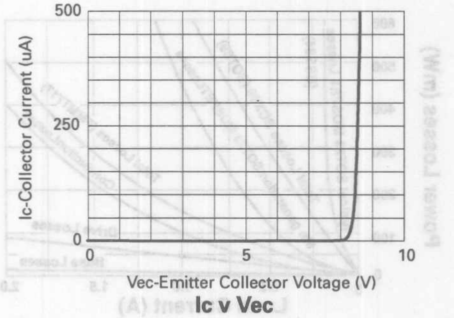


Figure 4
Reverse Blocking Capability of Bipolar Transistors.

8. Bipolar transistors can sometimes offer the useful feature of **inverse gain** or h_{FC}. This is particularly the case for the lower voltage variants, where the inverse gain can peak at between 33 to 50% of the peak forward gain. This is because the relatively highly doped collector region can also function well as an "emitter". For the Super-β parts this presents a peak inverse gain in the region of 100 to 300 typically. This feature can be useful for instance in positive line switching networks, where the selection of supply lines may effectively reverse the collector-emitter bias seen by the pass transistor. Another application benefit of importance is the possibility of conducting negative transients, as caused by external influences on the supply rail or inductive loads. This feature can be used to advantage in some circuits by allowing the omission of the collector-emitter diode that would otherwise be required, to prevent damage to the transistors emitter-base junction. Figure 5 shows the inverse gain characteristic for low collector currents, in the saturation region for the FMMT717 PNP SOT23 transistor.

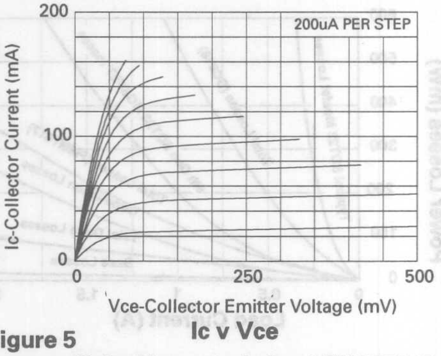


Figure 5
Inverse Gain Characteristic of FMMT717.

9 and 10. These last two factors are somewhat of an invention, but serve to demonstrate that when considered in terms of equivalent on-resistance (whether that is $R_{DS(on)}$ for a MOSFET, or $R_{CE(sat)}$ for a bipolar transistor) and most importantly **cost**, the bipolar option can be very attractive.

Figures 6 and 7 help to illustrate some of the points raised in 1 and 3 above, namely the silicon efficiency, on-state loss, and drive losses. These charts show the amount of power loss exhibited by a range of SOT23 and one SOT89 surface mount transistors as a function of load current. Figure 6 shows curves for NPN and N-Channel parts, and Figure 7 shows curves for PNP and P-Channel parts. These curves do **NOT** represent the minimum loss irrespective of package - they refer to industry standard and best-in-class SOT23 and SOT89 products only. It is noteworthy that the BCX69 part referenced is actually a SOT89 packaged transistor.

Key Parameters

The previous section has touched on some of the circuit parameters useful in comparing the two technology classes, when considering products for battery powered systems. Perhaps the most useful of these parameters is $R_{CE(sat)}$, this being the collector-emitter resistance, and is equivalent (in a datasheet sense) to $R_{DS(on)}$; the common MOSFET benchmark parameter. Table 2 shows a selection of the low voltage variant Zetex transistors specifically developed for battery system operation, and includes $R_{CE(sat)}$ figures for illustration purposes. As with any datasheet figure, point measurement values must be used as guidelines only. For application specifics, knowledge of the circuit operating conditions, datasheet test points and characterisation curves can usually lead to a credible interpolation.

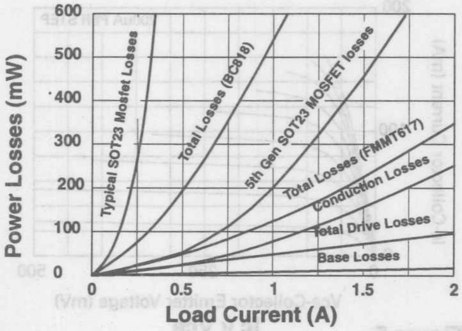


Figure 6
NPN/N-Channel Power Loss vs Load Current comparison.

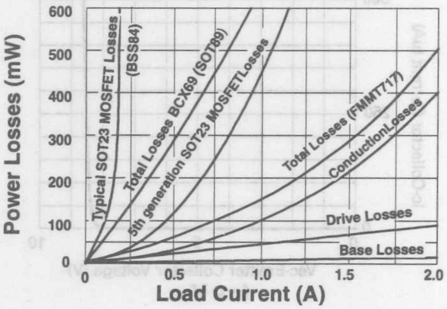


Figure 7
PNP/P-Channel Power Loss vs Load Current comparison.

Device (Note 1)	Polarity	I_C (DC)	I_{CM}	$V_{CE(sat)}$ (Note 2)	$R_{CE(sat)}$	h_{FE} (Note 3)
FMMT617	NPN	3	12	50mV @ 1A 150mV @ 3A	50mΩ	450 typ
FMMT618	NPN	2.5	6	50mV @ 1A 100mV @ 2A	50mΩ	450 typ
FMMT717	PNP	2.5	10	80mV @ 1A 140mV @ 2A	80mΩ	450 typ
FMMT718	PNP	1.5	6	110mV @ 1A 150mV @ 2A	55mΩ	450 typ
ZTX688B	NPN	3	10	50mV @ 1A 250mV @ 3A	50mΩ	750 typ
ZTX788B	PNP	3	8	50mV @ 0.5A 300mV @ 3A	100mΩ	650 typ
ZTX1048A	NPN	4	20	24mV @ 0.5A	48mΩ	450 typ
ZTX869	NPN	5	20	25mV @ 0.5A 180mV @ 5A	50mΩ	450 typ
ZTX949	PNP	4.5	20	40mV @ 0.5A 240mV @ 5A	80mΩ	200 typ

Table 2
Low Voltage transistors for Battery and Portable Systems.

Notes:

1. These represent a selection from some of the transistor families available from Zetex. Eg. the ZTX688B is a 12V 3A part from the ZTX688B-696B series, similarly the ZTX788B is part of the ZTX788B-795A series. Please note that surface mount equivalents are available for any ZTX (through-hole) pre-fixed part.

2. These $V_{CE(sat)}$ values are point measurements only, and depend on collector current, base drive level, and temperature. Please consult the appropriate datasheet for full DC characterisation. The corresponding values of $R_{CE(sat)}$ are shown for this point measurement, for application specific values consult the datasheet.

3. These values are guidance only, and are typical mid-band figures. Please refer to the appropriate datasheet. (Appendix A contains FMMT717 data).

The voltage rating parameters also require some examination. For bipolar parts the voltage rating most often quoted is the B_{VCEO} parameter, which is the collector-emitter breakdown voltage with the base terminal open-circuit. The primary breakdown of the epitaxial layer however is more closely represented by the B_{VCBO} or B_{VCES} parameters, and for many circuit topologies it is this rating that is of most relevance as the base is never open circuit; being driven actively, or tied to the appropriate rail by a resistor. For realistic comparisons with MOSFET devices, the primary value given by B_{VCBO} , B_{VCES}, or even B_{VCEV} should be considered if allowed by the circuit. This will probably lead to the selection of a lower voltage rated part than would otherwise be the case, leading to a lower $V_{CE(sat)}$ specification, and hence higher circuit performance. Please refer to Figure 8.

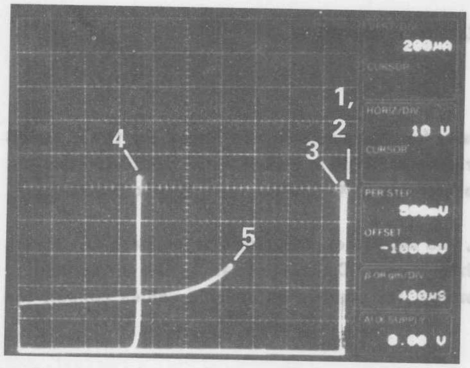


Figure 8
Voltage Breakdown Modes of ZTX849.

1 - BV_{CBO}, 2 - BV_{CES}, 3 - BV_{CEV}, 4 - BV_{CEO},
5 - BV_{CE} with $V_{BE} = 0.5V$
Scaling: 10V/div horiz.; 200µA/div vert.

Table 3 provides a guide to common bipolar/MOS terminology.

Bipolar	**MOSFET**
I_C (DC)	I_D (DC)
I_{CM}	I_{DM}
BV_{CES}	BV_{DSS}
$V_{BE(sat)}$	V_{GS}
I_{EBO}	I_{GSS}
I_{CES}	I_{DSS}
$R_{CE(sat)} = V_{CE(sat)}/IC$	$R_{DS(on)} = V_{DS(on)}/ID$
h_{FE}	g_{FS}
C_{ibo}	C_{iss}
C_{obo}	C_{oss}

Table 3
Bipolar/MOSFET Terminology.

Zetex datasheets for the very low $V_{CE(sat)}$, high current transistors include curves showing how $V_{CE(sat)}$ varies with both forced gain and temperature, as well as the more usual h_{FE} profile, $V_{BE(on)}$, $V_{BE(sat)}$ and safe operating area (SOA) charts. Appendix A presents an example of the characterisation available for many of the battery product targeted bipolar transistors: in this case the datasheet for the FMMT717 SOT23 PNP device. This is a 12V, 2.5A continuous high gain transistor that has been developed specifically for use as an high efficiency positive line switch for DC rail control, and high current DC-DC converters.

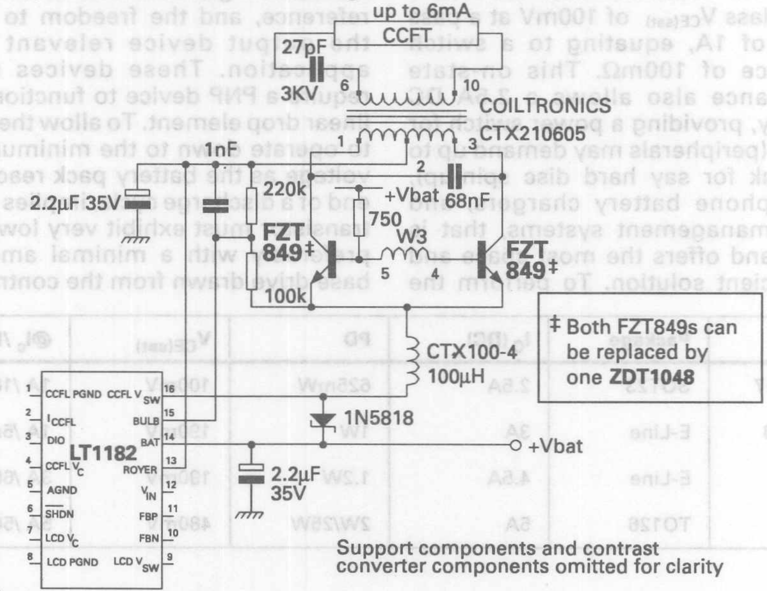
Applications

LCD Backlighting

Perhaps the most difficult power supply to effect within a laptop, and which has attracted much interest from many vendors, is the high voltage DC-AC inverter required by the fluorescent tube used to provide back/edge lighting for the LCD display. The tube expects a very high voltage to initiate conduction, perhaps 1kV, and several hundred volts during operation. This supply compliance must be effected with a very high degree of efficiency from the available energy source; - typically a ten cell NiCd or NiMH battery pack. The ZTX1048A series of transistors permit conversion efficiencies of over 90% providing significant increases in battery life, and therefore less re-charge cycles. The ZTX1048A and ZTX1049A devices have been developed specifically for the resonant push-pull (or Royer) inverter used almost exclusively for this purpose, and Zetex has already achieved many design-ins in this application. The '1048 and '1049 transistors are also available as an uncommitted dual device in the space saving SM-8 package as the ZDT1048 and ZDT1049.

[**Note:** The SM-8 package is an eight lead version of the popular SOT223 package and can thus yield a 50% space saving].

Figure 9 shows a floating lamp CCFL circuit developed by Linear Technology, using the LT1182 CCFL/LCD contrast dual switching regulator, and either a



‡ Both FZT849s can be replaced by one **ZDT1048**

Support components and contrast converter components omitted for clarity

Figure 9
Floating Tube CCFL Backlight Inverter (Linear Technology) using 2 x FZT849 or the ZDT1048 SM8 Dual transistor.

pair of FZT849s or a ZDT1048 SM-8 dual transistor to produce a 90% efficient inverter. Please refer to Linear Technology application note AN65 Oct 95, "A Fourth Generation of LCD Backlight Technology" by Jim Williams.

Power Supply Switching

Power supply switching for load management (such as peripheral control, transmit circuit blocks in handphones, and RAM back-up) should be considered carefully in terms of the DC and peak currents required; the allowable voltage drop across the switch element (the PCMCIA power switching specification for example, states a 5% maximum drop at a 1A output, from the nominal 5V or 3.3V rail); ease of drive and cost constraints. The FMMT717 is a SOT23 PNP transistor that exhibits a best in class $V_{CE(sat)}$ of 100mV at a pass current of 1A, equating to a switch resistance of 100mΩ. This on-state performance also allows a 2.5A DC capability, providing a power switch for PCMCIA (peripherals may demand up to 1.5A peak for say hard disc spin-up), mobile phone battery chargers, and battery management systems, that is reliable and offers the most space and cost efficient solution. To perform the same function with a P-Channel MOSFET would require a much larger die (and therefore package) and therefore an increase in price and weight for the user.

Analogue IC vendors manufacture microprocessor power supply supervisory devices, that monitor the state of memory supply rails and switch over the supplies as required. These devices often require a low loss PNP or P-Channel part as the pass element, and this function can usually be effected with much less cost by a suitable PNP transistor.

Voltage Regulation

Low drop out regulator controller ICs are now available that provide the user with the advantages of a monolithic voltage reference, and the freedom to specify the output device relevant to the application. These devices usually require a PNP device to function as the linear drop element. To allow the system to operate down to the minimum input voltage as the battery pack reaches the end of a discharge cycle, implies that the transistor must exhibit very low $V_{CE(sat)}$, preferably with a minimal amount of base drive drawn from the controller IC.

Device	Package	I_C (DC)	PD	$V_{CE(sat)}$	@I_c /I_b
FMMT717	SOT23	2.5A	625mW	100mV	1A /10mA
ZTX788B	E-Line	3A	1W	190mV	1A /5mA
ZTX949	E-Line	4.5A	1.2W	190mV	3A /60mA
ZBD949	TO126	5A	2W/25W	480mV	5A /50mA

Table 4
PNP Transistors for Low Dropout regulator Designs.

There are a number of Zetex parts ideally suited for this application, and Table 4 provides an overview of suitable parts. Figure 10 shows two typical LDO regulator circuits published by Linear Technology and MAXIM.

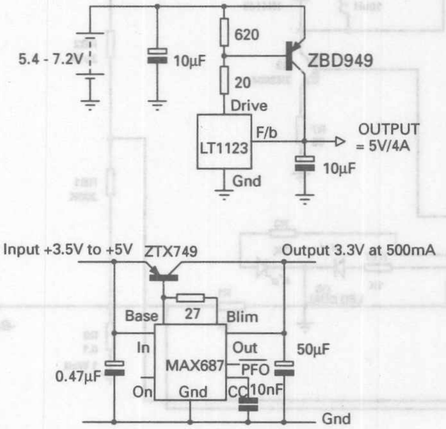


Figure 10
Low Drop Out Regulators - IC + Discrete Implementation.

DC-DC Converters and Fast Chargers

Battery charge management and fast charge circuit topologies using intelligent charge, voltage inflection, current, and cell temperature monitoring, frequently employ a DC-DC converter to effect an efficient charge transfer between the available source and the battery pack. This is usually a step down or Buck converter which dictates a fairly high operating frequency (to meet inductor size constraints), and is also subject to very tight cost control. To meet these requirements, Zetex have a range of high current, high gain transistors available in E-Line, SOT223 and TO126 that allow converter designs to 100kHz, and fast charge currents of up to 5A. Figure 11 shows a circuit designed by Benchmarq Microelectronics Inc., that uses the 3A rated ZTX788B for a low cost fast charger running at typically 80kHz and supplying a charging current of 2.3A. (This particular circuit being configured to charge two cells, and will accept an input voltage up to 15V). The circuit uses a turn-off circuit devised by Benchmarq to remove switching losses associated with bipolar transistor storage time and turn-off fall time, to allow the transistors to exhibit similar switching efficiences to large MOSFETs. This switch-off circuit is also shown in Figure 12. Q2 is driven by the PWM switching controller, and with the emitter resistor sets the base current for the high current PNP. These components are selected to ensure that on-state losses are acceptable for a given load condition without incurring excessive drive loss. When Q2 switches off, the inductor L1 rings, turning Q1 on hard. Q2 then performs active pull-up on the base of Q3 -the switching transistor. This method, and similar circuit techniques to remove base charge can be used to allow cost effective bipolar DC-DC converters.

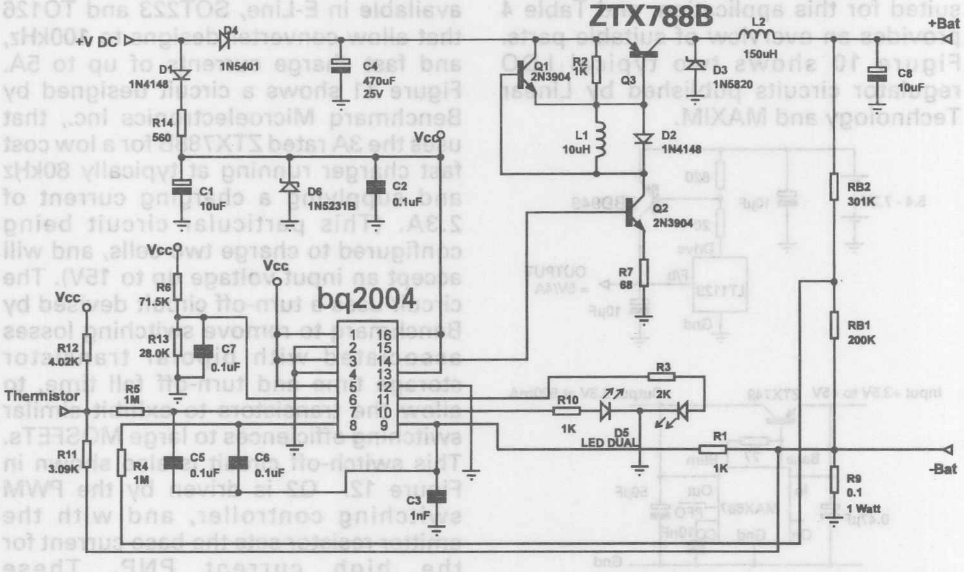


Figure 11
Fast Charge Circuit (Benchmarq Microelectronics) using the ZTX788B.

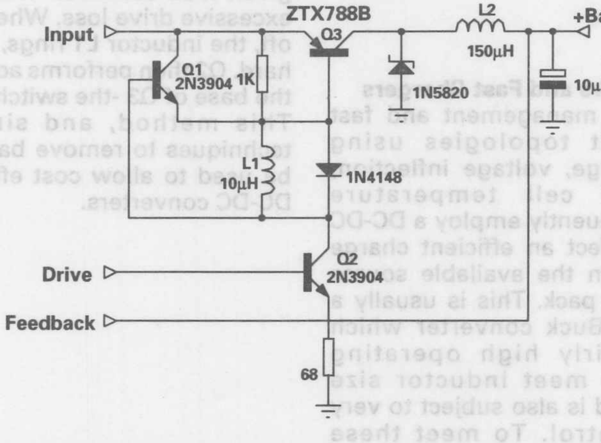


Figure 12
Turn-off circuit for Bipolar Transistors, allowing High Efficiency DC-DC Conversion at High Frequency.

APPENDIX A

FMMT717 datasheet including absolute maximum ratings, detail sheet and full DC characterisation.

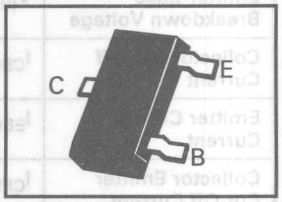
“SuperSOT”

FMMT717

SOT23 PNP SILICON POWER (SWITCHING) TRANSISTOR

FEATURES

- **625mW POWER DISSIPATION**
- **I_C CONT 2.5A**
- 10A PEAK PULSE CURRENT
- EXCELLENT H_{FE} CHARACTERISTICS UP TO 10A (PULSED)
- LOW SATURATION VOLTAGE



SOT23

COMPLEMENTARY TYPE – FMMT617
PARTMARKING DETAIL – 717

ABSOLUTE MAXIMUM RATINGS.

PARAMETER	SYMBOL	VALUE	UNIT
Collector-Base Voltage	V_{CBO}	-12	V
Collector-Emitter Voltage	V_{CEO}	-12	V
Emitter-Base Voltage	V_{EBO}	-5	V
Peak Pulse Current **	I_{CM}	-10	A
Continuous Collector Current	I_C	-2.5	A
Base Current	I_B	-500	mA
Power Dissipation at T_{amb}=25°C*	P_{tot}	-625	mW
Operating and Storage Temperature Range	T_j:T_{stg}	-55 to +150	°C

*Maximum power dissipation is calculated assuming that the device is mounted on a ceramic substrate measuring 15x15x0.6mm

**Measured under pulsed conditions. Pulse width=300µs. Duty cycle ≤ 2%

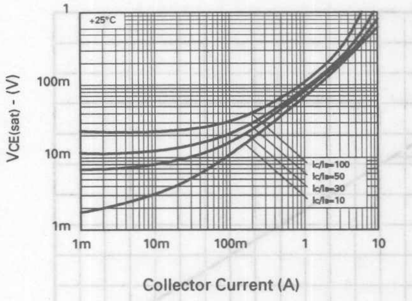
ELECTRICAL CHARACTERISTICS (at T_{amb} = 25°C unless otherwise stated).

PARAMETER	SYMBOL	MIN.	TYP.	MAX.	UNIT	CONDITIONS.
Collector-Base Breakdown Voltage	$V_{(BR)CBO}$	-12	-35		V	I_C=-100μA
Collector-Emitter Breakdown Voltage	$V_{(BR)CEO}$	-12	-25		V	I_C=-10mA*
Emitter-Base Breakdown Voltage	$V_{(BR)EBO}$	-5	-8.5		V	I_E=-100μA
Collector Cut-Off Current	I_{CBO}			-100	nA	V_{CB}=-10V
Emitter Cut-Off Current	I_{EBO}			-100	nA	V_{EB}=-4V
Collector Emitter Cut-Off Current	I_{CES}			-100	nA	V_{CES}=-10V
Collector-Emitter Saturation Voltage	$V_{CE(sat)}$		-10 -100 -110 -180	-17 -140 -170 -220	mV mV mV mV	I_C=-0.1A, I_B=-10mA* I_C=-1A, I_B=-10mA* I_C=-1.5A, I_B=-50mA* I_C=-2.5A, I_B=-50mA*
Base-Emitter Saturation Voltage	$V_{BE(sat)}$		-0.9	-1.0	V	I_C=-2.5A, I_B=-50mA*
Base-Emitter Turn-On Voltage	$V_{BE(on)}$		-0.8	-1.0	V	I_C=-2.5A, V_{CE}=-2V*
Static Forward Current Transfer Ratio	h_{FE}	300 300 180 60 45	475 450 275 100 70			I_C=-10mA, V_{CE}=-2V* I_C=-100mA, V_{CE}=-2V* I_C=-2.5A, V_{CE}=-2V* I_C=-8A, V_{CE}=-2V* I_C=-10A, V_{CE}=-2V*
Transition Frequency	f_T	80	110		MHz	I_C=-50mA, V_{CE}=-10V f=100MHz
Output Capacitance	C_{obo}		21	30	pF	V_{CB}=-10V, f=1MHz
Turn-On Time	$t_{(on)}$		70			V_{CC}=-6V, I_C=-2A I_{B1}=I_{B2}=50mA
Turn-Off Time	$t_{(off)}$		130			

*Measured under pulsed conditions. Pulse width=300μs. Duty cycle ≤ 2%

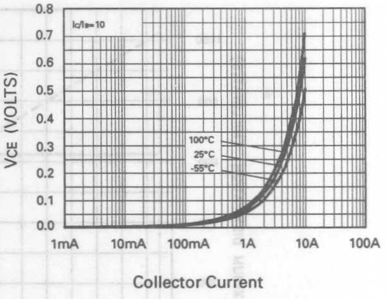
ZETEX

TYPICAL CHARACTERISTICS



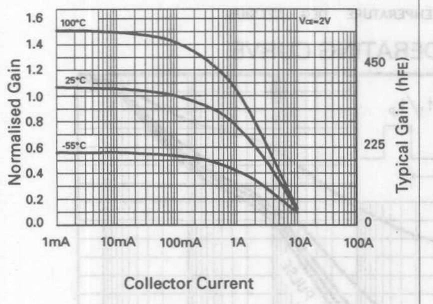
Collector Current (A)

VCE(SAT) vs IC



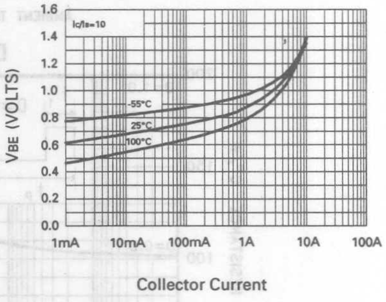
Collector Current

VCE(SAT) vs IC



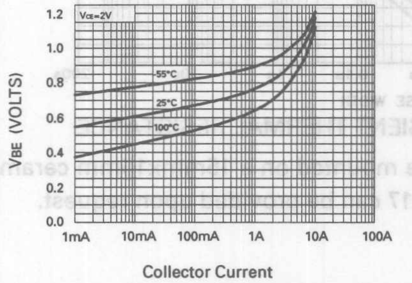
Collector Current

hFE vs IC



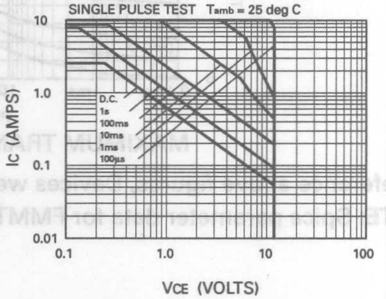
Collector Current

VBE(SAT) vs IC



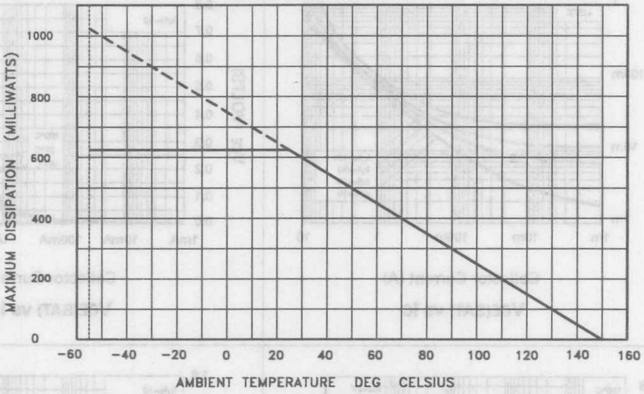
Collector Current

VBE(ON) vs IC

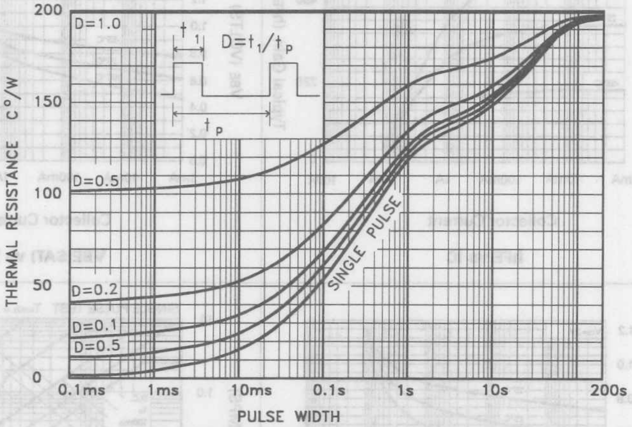


Safe Operating Area

THERMAL CHARACTERISTICS



DERATING CURVE



MAXIMUM TRANSIENT THERMAL RESISTANCE

* Reference above figures, Devices were mounted on a 15mmx15mm ceramic

NOTE: Spice parameter data for FMMT717 can be provided upon request.

Application Note 22
Issue 1 May 1996

High Frequency DC-DC Conversion using High Current Bipolar Transistors

400kHz Operation with Optimised Geometry Devices

Neil Chadderton
Dino Rosaldi

Introduction

DC-DC conversion is one of the fundamental circuit functions within the electronics industry and addresses a wide field of market sectors, applications and supply requirements. A general trend, (to pursue cost, size and reliability advantages) has been to reduce the size of the DC-DC converter systems, and as the inductive elements of such a system are quite often the bulkiest components, increases in switching frequency provide one method to allow this. Switching frequencies therefore have increased from tens of kHz to hundreds of kHz, which has forced a revision of the enabling switching devices.

It is an often assumed maxim that bipolar transistors are useful for DC-DC conversion functions up to perhaps 50kHz, and for service beyond this frequency, that MOSFETs provide the only solution. The purpose of this application note is to demonstrate that bipolar transistors possessing superior geometrys - thereby providing a higher current capability per die area, can and are operated to reasonably high switching frequencies with minimal loss. This often allows a more compact design (due to the higher silicon efficiency of bipolar technology) and lower cost.

Background

To obtain the most efficient performance from bipolar transistors at high switching frequencies, an appreciation of the basic switching mechanisms and base charge analysis is useful. Appendix A provides an introduction to bipolar transistor switching behaviour, and appendix B provides references for mathematical treatments.

There are various methods for increasing the maximum switching speed of bipolar transistors. Some of these rely on preventing saturation of the device, thus vastly reducing the stored charge, while other methods remove the charge at turn-off. Figures 1, 2 and 5 show a number of common speed-up networks. Figures 1a through 1d show various options for preventing saturation, and effectively reduce/limit base drive when the collector terminal has fallen to a specific level. (Figure 1a is often termed a Schottky transistor,

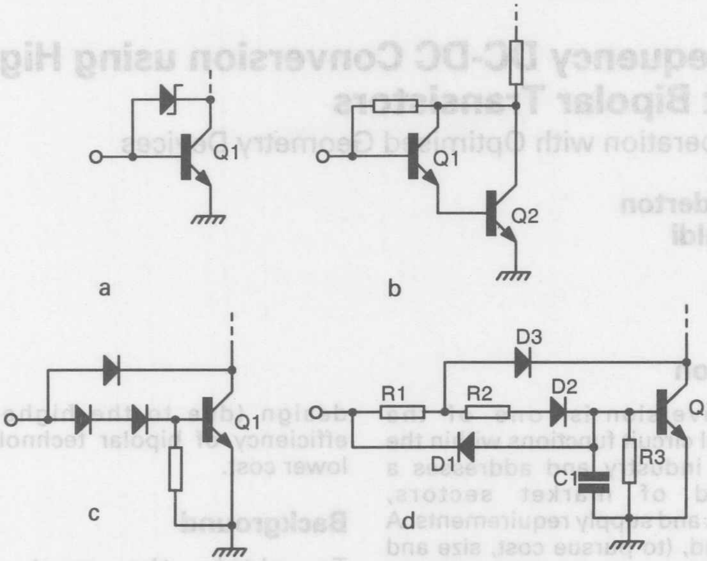


Figure 1.
Non-saturating Bipolar Transistor Speed-up Methods.

and is often used for IC transistors, and Figures 1c and 1d are variants of the so called Baker clamp circuit).

These methods (the Baker clamp being preferred for high power applications) of course do not allow the transistor's collector-emitter voltage to saturate to a very low level, and so the resultant on-state loss may be high and therefore prohibit the use of smaller packaged, though adequately current capable devices. Figures 2 and 5 show two methods that allow true saturation by permitting sufficient forward base drive, while removing charge quickly at turn-off. Design of such networks is non-critical, and by suitable choice of components allow high levels of base overdrive with no penalty to turn-off time duration.

Figure 2 shows a method of speeding up PNP devices to enable replacement of large P-Channel MOSFETs - this particular circuit being designed for fast charging of battery packs by Benchmarq Microelectronics. Figures 3 and 4 show the relevant waveforms, for a standard passive turn-off method (base-emitter pull-up resistor) and the speed-up circuit of Figure 2 - the combined storage and fall times being reduced from 1.2µs to 80ns. Importantly, the major switching loss contributor - the fall time, has been significantly reduced to 40ns, which is comparable to, or better than P-Channel MOSFET performance. This allows cost effective replacement of large packaged devices.

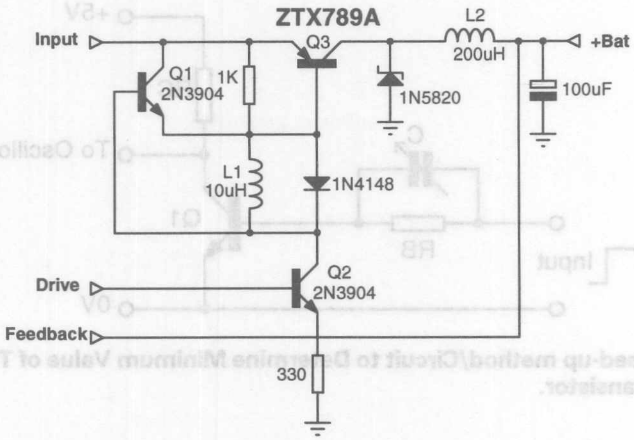


Figure 2.
PNP Transistor Speed-up Circuit to Allow Replacement of P-Channel MOSFETs within High Current Converters. (The circuit shown is suitable for output currents up to 1.5A; other variants are capable of operation to 5A output).

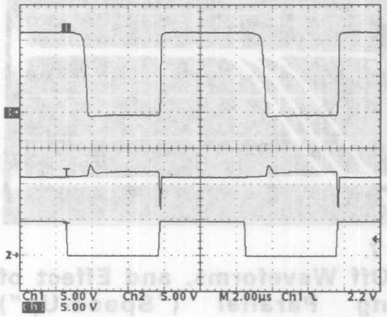


Figure 3.
Turn-off Waveforms for PNP Step-down Converter using Passive Turn-off.
Upper Trace - Collector-to-0V, 5V/div;
Middle Trace - PNP base-to-0V;
Lower Trace - PWM IC Output, 5v/div.
Timebase at 2µs/div.

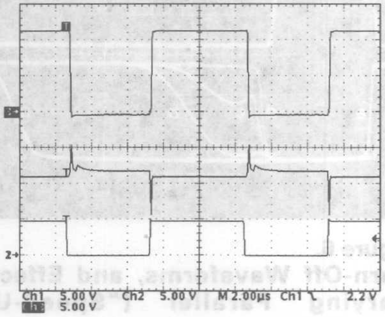


Figure 4.
Turn-off Waveforms for PNP Step-down Converter (of Figure 2) using Active Turn-off.
Upper Trace - Collector-to-0V, 5V/div;
Middle Trace - PNP base-to-0V;
Lower Trace - PWM IC Output, 5v/div.
Timebase at 2µs/div.

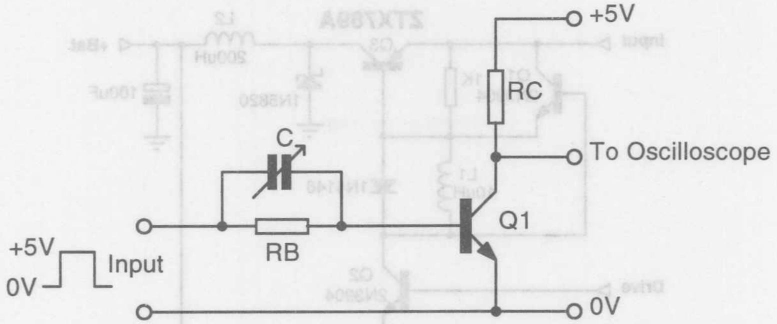


Figure 5.
Capacitive Speed-up method/Circuit to Determine Minimum Value of Turn-Off Charge of a Bipolar Transistor.

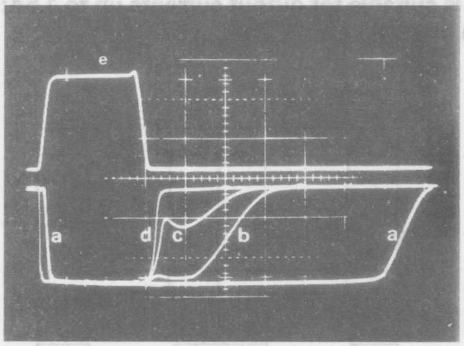


Figure 6.
Turn-Off Waveforms, and Effect of Varying Parallel ("Speed-Up") Capacitance.
Upper trace - Input waveform; lower traces - (a) C=0, (b) C=100pF, (c) C=120pF, (d) C=130pF (stored charge just removed).

Horizontal scale=200ns/div, Vertical scale=2V/div. ZTX300, I_B=1mA, I_C=10mA.

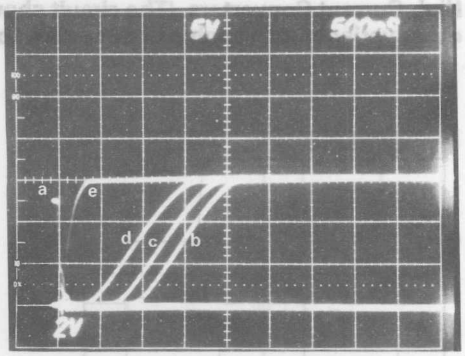


Figure 7.
Turn-Off Waveforms, and Effect of Varying Parallel ("Speed-Up") Capacitance.
Trace (a) - Input waveform; turn-off traces - (b) C=0, (c) C=470pF, (d) C=820pF, (e) C=1.5nF (stored charge removed).

Horizontal scale=500ns/div, Vertical scales: Input=2V/div; collector=5V/div. ZTX1048A, I_B=10mA, I_C=500mA.

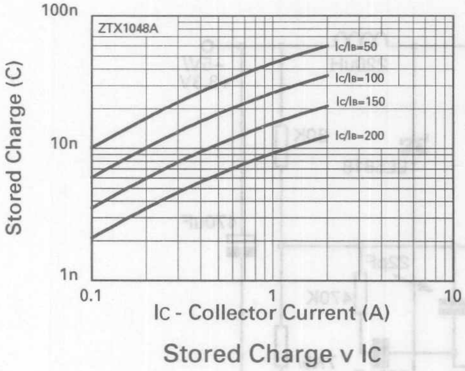


Figure 8.
Turn-off Charge Required as a Function of Collector Current and Forced Gain, for the ZTX1048A Bipolar Transistor.

Figure 5 shows the standard speed-up capacitance method, although in practice a finished design would use a fixed value capacitor. A similar circuit can also be used to determine the capacitance required for a particular bias condition. In practise the variable capacitance (or value of parallel capacitance) is adjusted such that the sum of stored charge and junction capacitance charge is just removed - allowing for device variation, temperature and bias tolerance.

The oscillographs in Figures 6 and 7 show the effect of increasing the capacitor value from zero (Eg. open circuit) to a value adequate to neutralise the stored charge. Figure 6 is for a small signal device, while Figure 7 is for the ZTX1048A - a transistor utilising the Zetex Matrix geometry, and a Super-β emitter process to produce a 4A DC rated part in the TO92 style E-Line package. Figure 8 summarises a set of such measurements for the ZTX1048A - for example, at a collector current of 1A and a forced gain of 200, a turn-off charge of 900pC is required to neutralise stored charge and eliminate storage time effects.

[**Note 1**: Appendix A also includes some ancillary material on the minimum amount of trigger charge necessary for pulse circuits].

Power Conversion Circuits

To demonstrate the performance advantages possible when the bipolar transistor's turn-off charge is addressed, this section considers such modifications to a basic step-down DC-DC converter. The circuit shown in Figure 9 was used to provide a means of evaluation, and is of fairly standard implementation, apart from the choice of the FMMT718 SuperSOT SOT23 PNP transistor. This circuit, with minimal design optimisation, can produce the efficiency against load current characteristic shown in Figure 10. This chart also shows how the bias conditions for the pass device can be modified to increase the current capability of the circuit, albeit with some compromise to conversion efficiency at lower load currents. Curves 1, 2 and 3 illustrate this effect for base currents of 9.4mA, 43mA and 170mA respectively. Figure 11 shows how the efficiency varies with input voltage for the I_B=43mA option. Higher output current designs are possible with larger die transistors from the Zetex range, such as the ZTX788B, ZTX789A, ZTX790A, ZTX948 and ZTX949. For a comprehensive listing please refer to Semiconductor Data Books one and two.

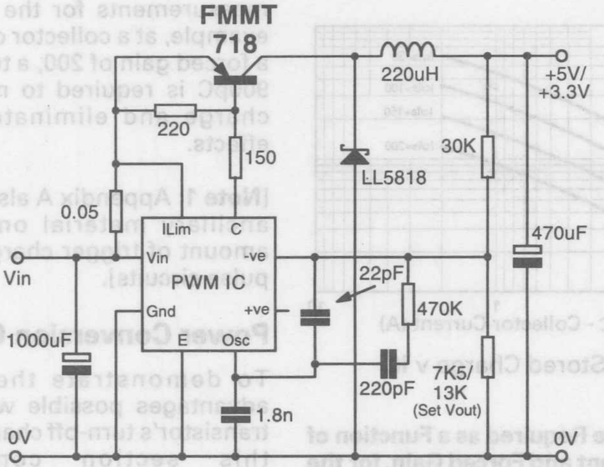
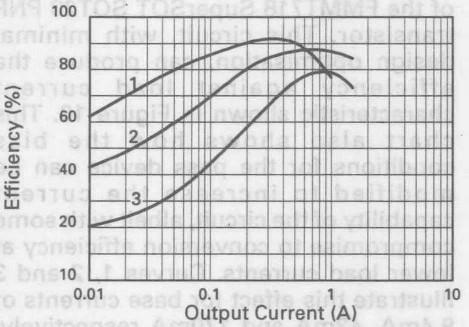
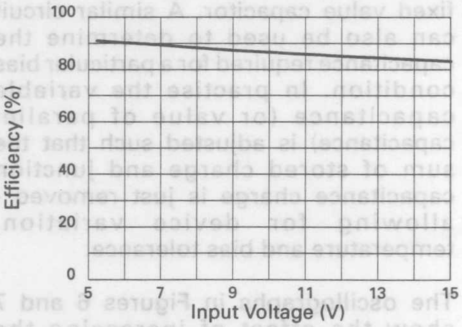


Figure 9.
Basic Step-Down DC-DC Converter using the LM3578 and the FMMT718.
(With values shown, F_{op}=50kHz, I_B=43mA, peak efficiency=88%).



Efficiency v Output Current

Figure 10.
Efficiency against Load Current for the Converter of Figure 9, with I_B as a parameter.
Curve 1 - I_B=9.4mA; Curve 2 - I_B=43mA; Curve 3 - I_B=170mA. F_{op}=50kHz; V_{in}=7V; V_{out}=5V.



Efficiency v Input Voltage

Figure 11.
Efficiency against Input Voltage for the Converter of Figure 9. (I_B=43mA; I_{out}=1A).

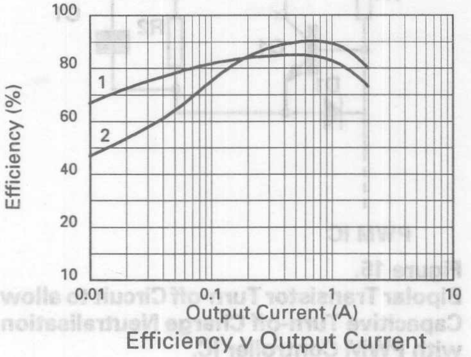


Figure 12.
Efficiency against Load Current for the Converter of Figure 9.
I_B=43mA; F_{op}=90kHz; V_{in}=7V; V_{out}=5V.
Effect of Variation of Inductor Value.
Curve 1 - 25µH, Curve 2 - 180µH.

Optimisation, and some scope for higher frequency operation is possible with this basic converter. Figure 12 shows the effect of varying the inductors value, with the switching frequency set to 90kHz. Turn-on time was recorded at 55ns, and turn-off at 1.5µs.

To investigate options for higher switching frequencies requires a change to the PWM controller IC. Figure 13 shows a circuit based on the TI5001 device. This IC is capable of operation up to a switching frequency of 400kHz, and can sink a maximum of 20mA into the VO pin.

The initial circuit was configured to operate at 150kHz, with the FMMT718 base current set to 9mA by a 560Ω base resistor (R1) - effectively a forced gain of 222 at 2A. Turn-on and turn-off times were measured at 200ns and 1.44µs

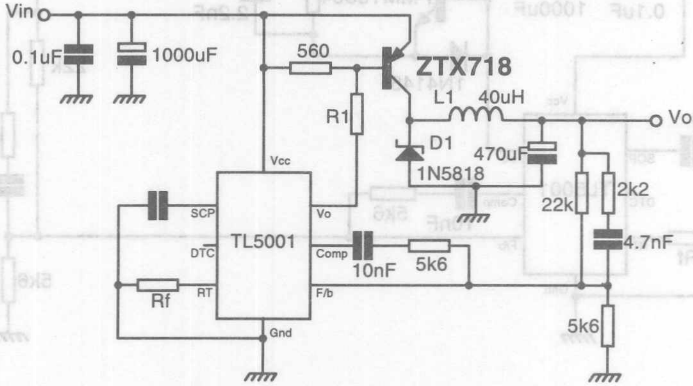


Figure 13.
Basic Step-Down DC-DC Converter using the TI5001 and the FMMT718. (With values shown, F_{op}=150kHz, I_B=9mA, peak efficiency=81%).

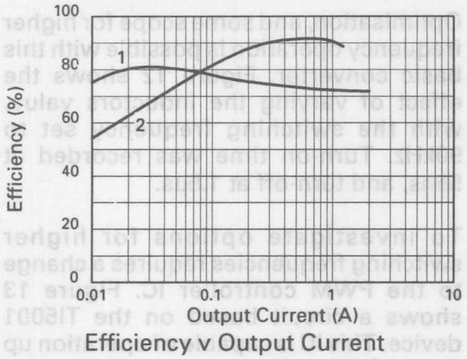


Figure 14.
Efficiency against Load Current for the Converters of Figure 13 and 16.

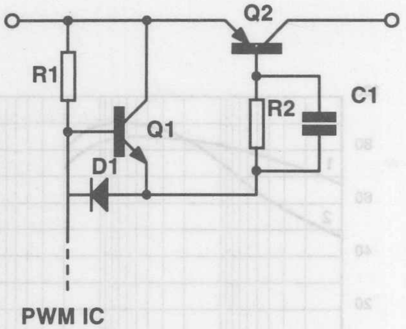


Figure 15.
Bipolar Transistor Turn-off Circuit to allow Capacitive Turn-off Charge Neutralisation with PWM Controller IC.

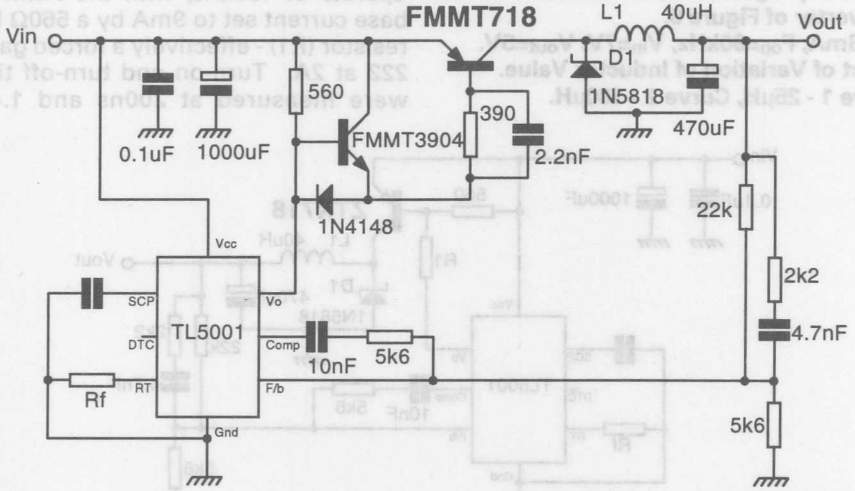


Figure 16.
Basic Step-down DC-DC Converter using the Tl5001/FMMT718 Combination with Capacitive Turn-off Circuit.

ZETEX

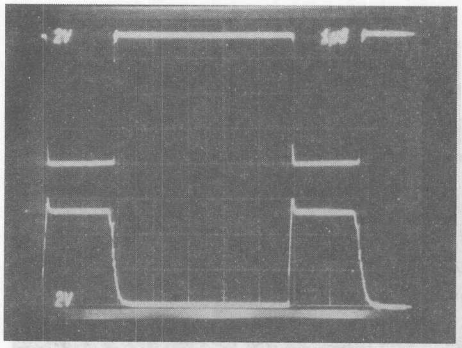


Figure 17.
Switching Waveforms for the circuit of Figure 16, operating at 150kHz. Upper trace; 2V/div, Collector-to-0V. Lower trace; 2V/div, Base-to-0V. V_{in}=7V, V_{out}=5V, I_L=220mA.

respectively. The chart shown in Figure 14 (curve 1) illustrates the resultant efficiency versus load profile. The circuit's conversion efficiency peaks at 81% at 23mA output, but falls thereafter to 71% at 2A, due mainly to switching losses, though in some part to increasing $V_{CE(sat)}$ at higher currents. The latter factor could be addressed in part by considering the load requirements, and optimising for the relevant condition, but the switching losses would remain.

By considering the base turn-off charge and effecting a suitable turn-off circuit (Figure 15) the circuit shown in Figure 16 was produced. This shows a significant improvement in conversion efficiency at medium to higher currents; as shown in

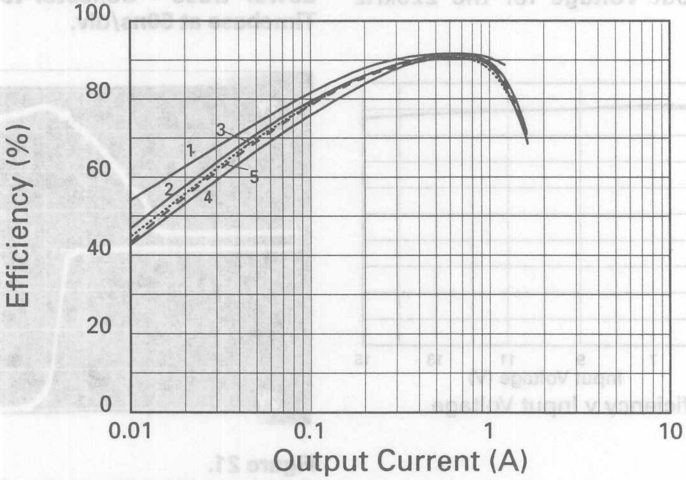
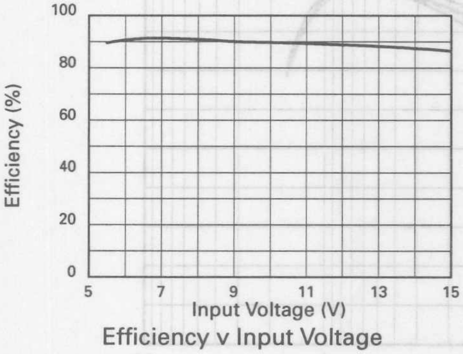


Figure 18.
Efficiency against Load Current for the Converter of Figure 16. Curve 1 - F_{op}=150kHz; curve 2 - 220kHz; curve 3 - 300kHz; curve 4 - 400kHz. I_B=10mA. Curve 5 - 400kHz and I_B=5mA. V_{in}=7V, V_{out}=5V.

Figure 14, curve 2. This shows a peak efficiency of 92%. Figure 17 shows the switching waveforms, including the collector-to-0V waveform - note the rapid turn-off edge; this was measured to be 25ns. The efficiency at lower currents has of course reduced, due to the extra current taken by the turn-off circuit. This efficiency profile can be modified for specific applications by sacrificing high current efficiency in favour of low current performance or vice-versa.

Further modifications were effected to assess performance at still higher switching frequencies. Figure 18 shows the efficiency/load current profiles for switching frequencies from 150kHz to 400kHz. Figure 19 shows the efficiency against input voltage for the 220kHz



Efficiency v Input Voltage

Figure 19.
Efficiency against Input Voltage for the Converter of Figure 16.
F_{op}=220kHz; V_{out}=5V; load current=1A.

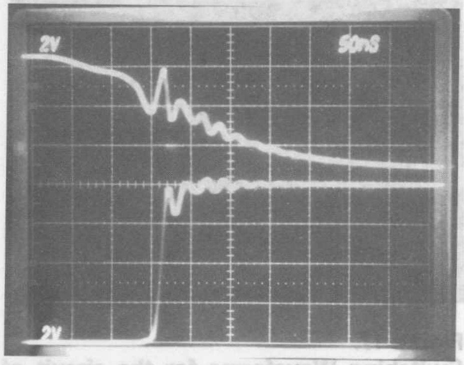


Figure 20.
Switching Waveforms for the Converter of Figure 16 operating at 400kHz.
Turn-on edges. Risetime= 20ns.
Upper trace - Base-to-0V, 2V/div.
Lower trace - Collector-to-0V, 2V/div.
Timebase at 50ns/div.

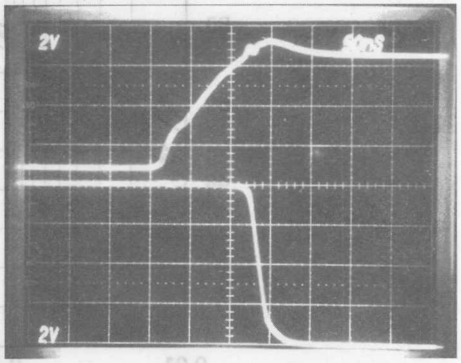


Figure 21.
Switching Waveforms for the Converter of Figure 16 operating at 400kHz.
Turn-off edges. Falltime=30ns.
Upper trace - Base-to-0V, 2V/div.
Lower trace - Collector-to-0V, 2V/div.
Timebase at 50ns/div.

version - the curve varying little over the measured range. Figure 18 curve 5, is again for a switching frequency of 400kHz, but with lower base drive (R_B=680Ω) and a reduction to 1.5nF for the turn-off capacitor. The oscillographs shown in Figures 20 and 21 show the turn-on and turn-off switching waveforms for the 400kHz version. These show collector rise and fall times of 20ns and 30ns respectively.

Conclusions

This application note has demonstrated that with due attention to the base and collector charge phenomena in bipolar transistors, the operating switching frequency of those parts can be extended well beyond the currently accepted notional maximum of 40kHz, to several hundred kHz. Various speed-up methods have been summarised, and examples of those particularly suitable and complementary to high current capable transistors examined, and circuit examples presented.

Appendix A

Bipolar Transistor Switching Behaviour

In most switching circuits trigger capacitors are required to differentiate pulses and provide DC isolation between stages, and speed-up capacitors are needed to neutralise the stored charge and junction capacitances of the transistors. In most cases it is sufficient to choose values small enough not to interfere with the operation of the circuit at its maximum frequency. These capacitors do however have minimum values depending on the type of transistor used, the base and collector currents, and the amplitude of the waveforms involved.

In order to understand the requirement for speed-up capacitors, consider the simple inverter circuit shown in Figure 22, and the resultant switching waveforms shown schematically in Figure 23.

Initially Q1 is cut off with its base-emitter (b-e) junction reverse biased to -1.67V and its collector-base (c-b) reverse biased by 6.67V. On the rising edge of the input pulse the potential at point A (V_{BE}) rises exponentially as the junction capacitances of the transistor charge, until the V_{BE} reaches 0.6-0.7V and the b-e junction starts to conduct significant current. The time before this occurs from the start of the input pulse is termed the delay time, t_d. As the base current (I_B) flows, charged carriers accumulate in the base region, and the collector current increases in proportion until it is

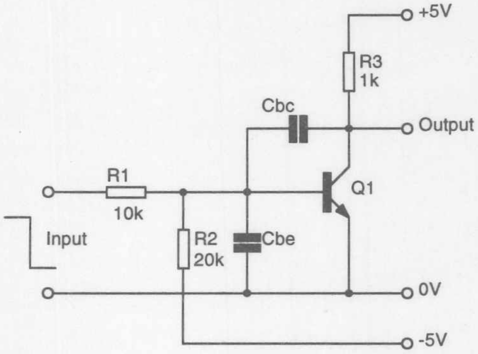


Figure 22.
Bipolar Transistor Switching Circuit for Illustration of Basic Switching Behaviour.

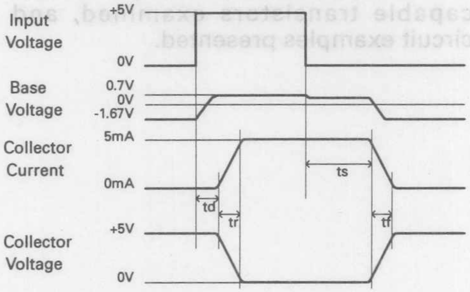


Figure 23.
Bipolar Transistor Switching Waveforms for the Inverter Circuit of Figure 22.

limited by the collector load resistance R3. The time taken for the collector current to rise from 10% to 90% of its final value is the rise time, t_r. As the collector voltage approaches zero, the c-b junction becomes forward biased and conducts the excess base current. The base now starts acting as an emitter and begins to inject charge carriers into the collector. This causes a considerable accumulation of charge in the collector region which must be removed, either by recombination or by reverse base current, before the device can begin to turn off. The time taken for this excess charge to be removed is called the storage time (t_s), and during this time the transistor remains saturated. At the end of the storage time all excess charge carriers have been removed and the base charge is simply that required to maintain collector current. As this charge is reduced further the collector current falls in sympathy until the device is cut off. The time taken for the collector current to fall from 90% to 10% is called the fall time, t_f.

Due to the low leakage of silicon transistors, it is not usually necessary to reverse bias the base to ensure DC stability. This does tend to reduce the delay time, but increases the storage and fall times as the stored charge can only be dissipated by recombination instead of by the reverse current provided by R2. This can be overcome by including a capacitor in parallel with the base drive resistor, of such value that the charge stored on it is sufficient to neutralise the total charge in the transistor. This value will be dependent on the transistor type, the base and collector currents and the amplitude of the input or driving pulse. For practical applications it is necessary to determine the required capacitor values by measurement, preferably using a worst case version of the circuit. As a guide, minimum values of turn-off charge have been provided for small signal, switching and high current low $V_{CE(sat)}$ transistors, and are reproduced for various DC conditions within Appendix C. These charts were produced using test circuits similar to that shown in Figure 5. As carrier diffusion coefficients and recombination are temperature dependent, then storage time will also possess this dependence to some degree. Therefore in some circumstances allowance must be made, and component values adjusted accordingly.

The mount of trigger charge necessary for pulse applications can be considered in a similar manner, and for obtaining approximate values the circuit shown in Figure 24 can be employed. The values obtained are higher than for the "speed-up" case as the trigger capacitor not only has to remove the stored charge but also has to overcome the DC bias provided by R_B. In a practical circuit, there may well be additional components, which will absorb some fraction of the trigger charge. Therefore it will usually be necessary to use a higher value of capacitance than measurement (or provided curves) would suggest - it is advised to closely breadboard/model the actual circuit as closely as possible with worst case component values. Figure 25 shows an example oscillograph for the ZTX300 small signal device, and Figure 26 presents curves for the minimum trigger

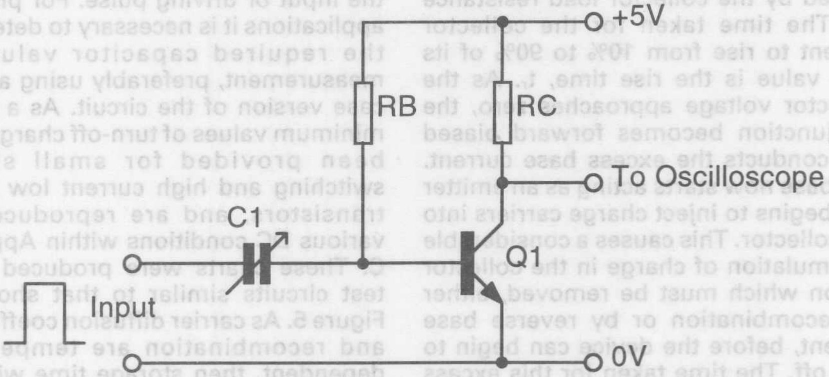


Figure 24.
Circuit to Determine Minimum Value of Trigger Charge for Bipolar Transistor.

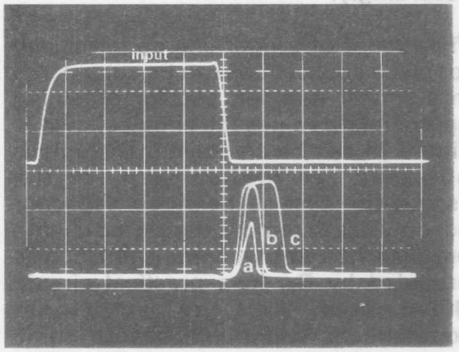
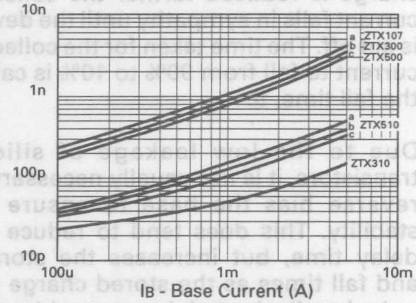


Figure 25.
Trigger Waveforms and Effect of Varying Trigger Capacitance. Upper trace - Input waveform; lower traces - output at collector, (a) Variable capacitor "C1" below critical value, (b) "C1" at critical value with collector voltage rising to 90% of final value, (c) "C1" above critical value. Horizontal scale=200ns/div, Vertical scale=2V/div. ZTX300, I_B=1mA, I_C=10mA.

charge necessary for small signal types ZTX107, ZTX300 and ZTX500, and switching transistors ZTX310 and ZTX510.



Min. Trigger Charge v IB for Small Signal and Switching Transistors

Figure 26.
Minimum Trigger Charge for Small Signal and Switching Transistors. Forced Gains (I_C/I_B) of : a)10; b)20; c)40.
No significant difference observed for different forced gains on the ZTX310 series.

Appendix B
References

(References 1, 2 and 3 contain mathematical treatment of charge analysis with respect to switching characteristics of bipolar transistors - similar background can be found in most semiconductor physics books).

1. Switching Transistor Handbook, chapter five - "Transient Characteristics of Transistors" . Motorola Inc., Technical Editor William D. Roehr.

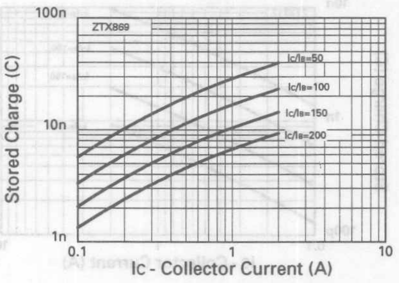
2. Pulse, Digital and Switching Waveforms, Chapter 20 - "Transient Switching Characteristics", Millman and Taub. McGraw-Hill Book Company.

3. Semiconductor Device Technology, section 2.2.26 BJT Modelling - Physical BJT Models, subsection 2) Charge Control BJT Model. Malcolm E. Goodge. MacMillan Press.

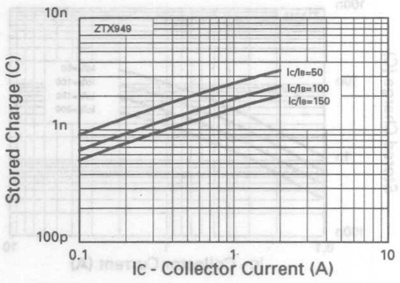
4. Switching and Linear Power Supply, Power Converter Design. Abraham I. Pressman. Hayden Press.

Appendix C

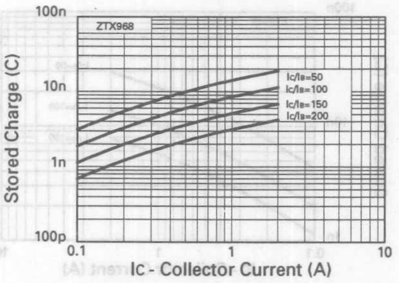
Characterisation Charts showing typical stored charge as a function of load current and bias level.



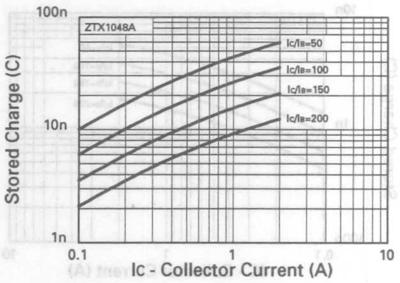
Stored Charge v Ic for ZTX869



Stored Charge v Ic for ZTX949



Stored Charge v Ic for ZTX968

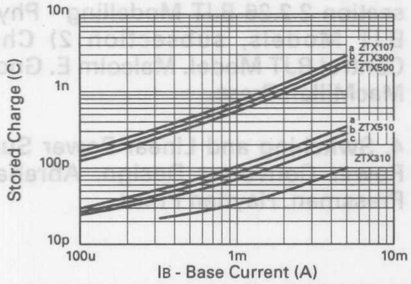


Stored Charge v Ic for ZTX1048A

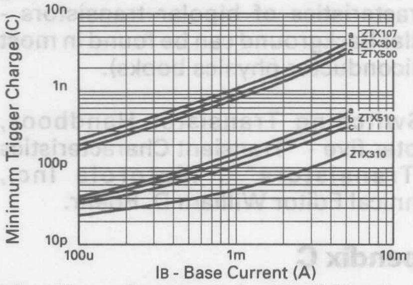
Appendix C (continued)

Forced gains (Ic/IB) of a:)10, b)20, C) 40.

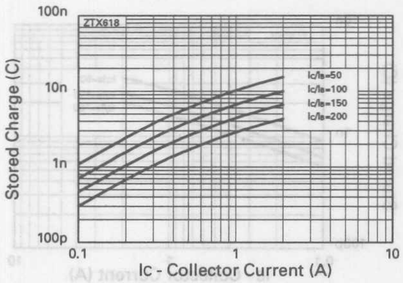
No significant difference observed for the different forced gains for the ZTX310 series



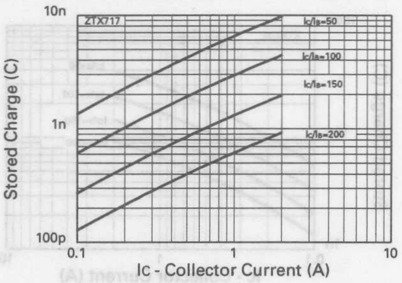
Stored Charge v IB for Small Signal and Switching Transistors



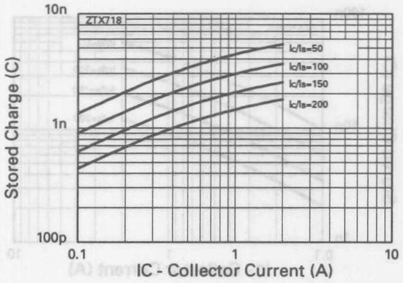
Min. Trigger Charge v IB for Small Signal and Switching Transistors



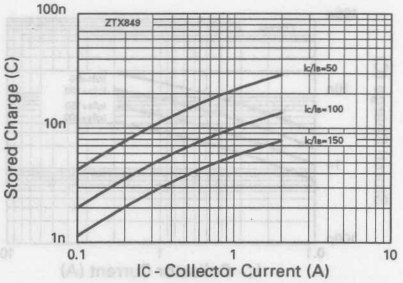
Stored Charge v Ic for ZTX618



Stored Charge v Ic for ZTX717



Stored Charge v Ic for ZTX718



Stored Charge v Ic for ZTX849

ZETEX

Application Note 23
Issue 2 March 1996

Zetex SPICE Models

Understanding Model Parameters and Applications Limitations

Neil Chadderton

Introduction

SPICE was originally developed as a simulation tool for Integrated Circuit design (SPICE being an acronym for Simulation Program with Integrated Circuit Emphasis) by the University of California in the 1970s. It was quickly improved into the SPICE2 version from which all commercial "Spice" programs are derived. It has since been enhanced (in terms of faster more robust algorithms and graphics handling capabilities for circuit input, and ease of analysing simulation output) and marketed by a number of companies for use within the PC environment - commercial versions being PSPICE, HSPICE, IsSPICE, SPICEAGE, Microcap etc.

Though originally intended for IC design, the availability of low cost computing, and the push towards robust design has introduced many other circuit and system designers to the advantages offered by analog circuit simulation. This has led to the requirement for device models for the active components under consideration, and so now many semiconductor companies provide appropriately targeted SPICE models as part of the technical support function.

These models can be extremely useful when used as a design tool, but care must be exercised. Any simulation software text will warn against sole reliance on the software's predictions. Models are (by definition) a compromise, and are essentially based on a device's common features. An appreciation of the model derivation, the model parameters, and it's inherent limitations should be sought, and would assist in the interpretation of simulation results and their application to the real world.

Zetex have created SPICE models for a range of semiconductor components. Many of these models are for the higher performance Bipolar and MOSFET transistors, but models for RF transistors, variable capacitance diodes, switching diodes and standard small signal parts are also available. These models are available through any Zetex sales office or agent.

Appendix A includes a printout of the introductory text file included in version 2 of the Zetex SPICE models disc. This file provides some background to how the model files are organised, and a brief overview of the model parameters for each type of device model.

This application note is a guide to the understanding and use of Zetex SPICE models. It includes sections on how some of these models are derived - the measurements/optimisation necessary, how to customise models for those cases where a model is not available, and the limitations to be aware of.

Measurement of Model Parameters

The bipolar transistor circuit model used by the SPICE software is a modified version of the Gummel-Poon model formulated in 1970. The schematic of this model is shown in Figure 1. For a comprehensive description of the model, including physical definitions of the model components, please refer to Appendix B, references 1 and 2.

Figure 2 shows a typical model for a Zetex bipolar transistor.

Lines beginning with an asterisk indicate a non-executable comment; so device details, date of creation, line spaces, and copyright messages all start with this symbol.

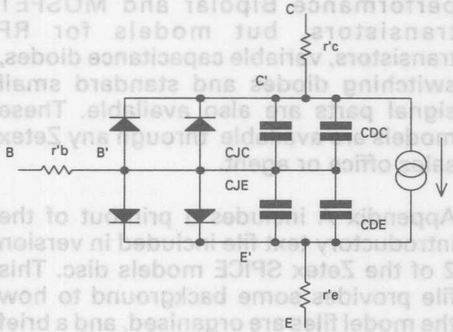


Figure 1
Gummel-Poon Bipolar Transistor Model.

The line beginning ".MODEL ...." uses a standard SPICE command that defines a model to the software. Following this command are the device name, it's polarity, and a list of the model parameters - the "+" sign being used for concatenation to the original command. The text file reproduced in appendix A gives some brief details as to the effects each parameter has on the model's behaviour.

The measurements necessary to derive the bipolar transistor model parameter values can conveniently be separated into dc and ac parameters as shown in table 1. The dc parameters are further separated into those for forward and reverse operation. The parasitic resistance components, and the saturation current IS being common for both forward and reverse modes. The descriptions of the forward parameters given below is equally applicable to the reverse parameters. (Appendix B reference 3).

dc		ac
forward	reverse	
IS		CJC
NF	NR	CJE
BF	BR	MJC
IKF	IKR	VJC
VAF	VAR	MJE
ISE	ISC	VJE
NE	NC	TF
RB		TR
RE		
RC		

Table 1
Bipolar Transistor Model Parameters.

```
*ZETEX     ZTX688B     Spice model     Last revision 8/11/90
*
.MODEL ZTX688B NPN IS = 1.09E-12 NF = 0.9935 BF = 1180 IKF= 5.2 VAF= 25
+ ISE=1.3E-13 NE =1.35 NR =0.992 BR =790 IKR=.5 VAR=5 ISC=0.174E-12
+ NC =1.399 RB = .3 RE =.036 RC = .034 CJC= 104E-12 MJC= .29 VJC= .46
+ CJE=280E-12 TF = .93E-9 TR =1.05E-9
*
```

Figure 2
General Format of Zetex (Bipolar Transistor) SPICE Models.

An explanation of the function of the dc parameters is most easily accomplished with reference to a chart - sometimes referred to as the "Gummel Plot". Figure 3 shows a Gummel plot, which illustrates the variation in collector and base currents with base-emitter voltage - a requirement being that the device is on the edge of saturation, satisfied by $V_{BC} = 0$, (or $V_{CE} = V_{BE}$). The collector current I_C (ideally) follows the Shockley equation that defines the current through a P-N junction:

$$I = I_s\,[e^{\frac{qV_f}{KT}} - 1]$$

where

I_S - the transistor saturation current and is dependent on the size of the base-emitter area

k - Boltzmann's constant=1.38 x 10^{-23}J/K

q - electronic charge=1.602 x 10^{-19}C

T - temperature in Kelvin, (300K is a common assumption)

V_F - is the voltage across the junction.

With reference to the chart, IS is the y-axis intercept of the collector current curve at $V_{BE}= 0$, and NF defines the slope of the line (referred to q/kT), and within SPICE defaults to unity. At high currents, the collector current differs from that predicted by the equation due to resistances, and high level injection effects. These effects are accommodated within SPICE by allowing base-emitter de-biasing by the parasitic emitter resistance (or collector resistance for the reverse mode), and by the collector "knee" current parameter, IKF. (IKF is defined as the collector current at which the current gain= BF/2, see below).

The base current (I_B) plot would ideally be parallel to the collector current, the plot separation being IS/BF, however there are a number of physical effects that are addressed by other parameters.

At low values of collector current, surface leakage and recombination effects introduce an additional slope to the plot of I_B, causing the line to intercept the y axis at a higher position than would be the case if I_B was always a fixed fraction of I_C. These effects are described by two additional SPICE parameters; ISE - the y axis intercept for the I_B curve at V_{BE}=0, and NE - representing the slope of the affected region. These parameters introduce a fall in h_{FE} at low collector current. (Appendix B, reference 4 gives some useful illustrations on how these parameters affect transistor curves).

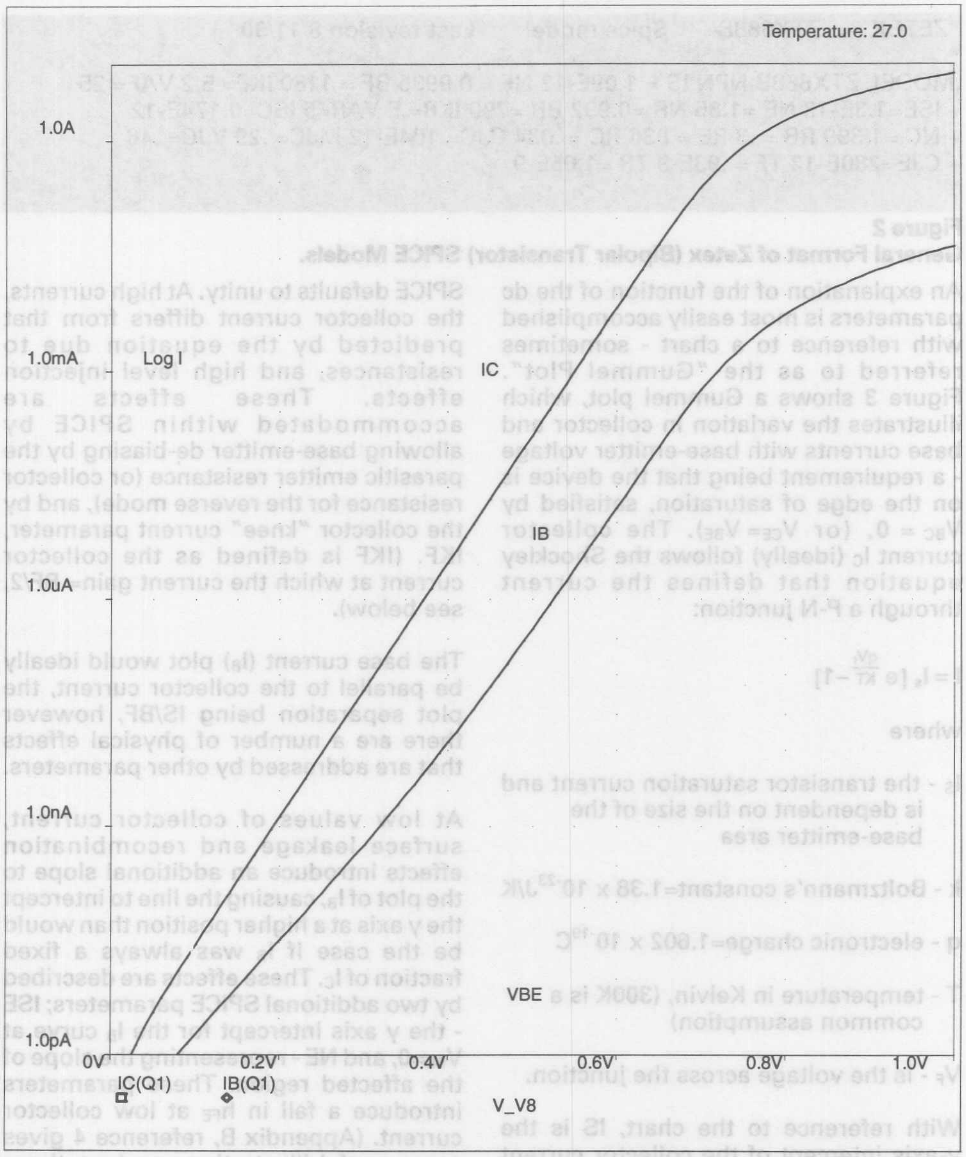


Figure 3
I_C and I_B Vs V_{BE} Transconductance Chart - "Gummel Plot" (FMMT617). (Note 1).

The Gummel Plot is produced either by careful curve tracer measurement, and/or by an arrangement of Source-Measure Units. The latter being particularly effective for low current measurements, and the former for the high current range. If more than one measurement system is used to produce Gummel data, then care must be exercised to ensure that temperature and accuracy differences do not produce offsets. The Gummel Plot measurement is usually automated, and the data collected into a PC where a logarithmic regression routine is used to determine IS, NF, ISE and NE.

The slope of the I_C versus V_{CE} output characteristics, (hoe in h-parameter parlance) is due to base width modulation effects as described by J.M. Early - and sometimes termed the Early effect, (please refer to Figure 4). The *Early voltage* is the point on the x axis at which the extrapolated curves would appear to intercept. This voltage (which possesses no sign) is known as VAF to the SPICE software, and allows the SPICE equations to predict the I_C value from the basic Shockley equation, at any value of V_{CE}. This allows the model to behave as expected in the linear region.

The VAF (VAR) parameters are determined by measuring I_C (I_E) at several values of V_{CE} (V_{EC}), and using linear regression to find the voltage axis intercept. A range of base currents are used, such that the resultant collector currents represent the usual operating range.

The ac parameters are the collector-base and base-emitter capacitances, and the forward and reverse base transition time parameters. The capacitance parameters CJ, MJ, and VJ are determined by measuring the capacitance against reverse voltage (known as C-V data) for each junction, and then either using an iterative routine based on the equation below, or the P-N junction capacitance entry screen within the PSPICE PARTS package, or similar parameter extraction program.

$$C = \frac{CJ0}{\left(1 - \frac{V_R}{\varphi}\right)^M}$$

where

CJ0 - is the zero bias value

φ- is the junction barrier potential (known as VJC or VJE to SPICE)

M - is the grading coefficient (known as MJC or MJE to SPICE).

V_R - is the applied reverse bias.

The transit time parameters can be found by an iterative approach based on a measurement of the F_T profile (transition frequency against collector current), and switching times for a particular range of conditions. The forward transit time TF is adjusted such that simulations of small signal RF gain measurements concur with F_T measurements (Appendix B reference 5) and turn-on times. The reverse transit time TR is determined by consideration of turn-off times, particularly bipolar storage time, ts.

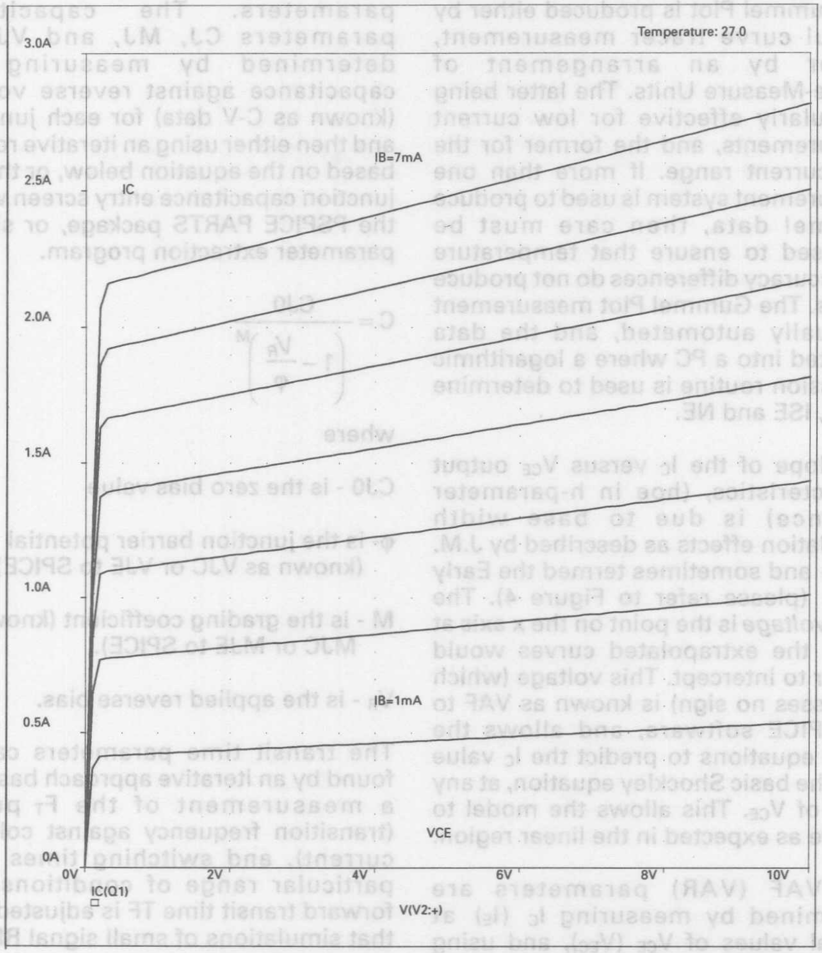


Figure 4
I_C Vs V_{CE} Output Characteristics Plot, Illustrating the "Early Effect" (FMMT617). (Note 1).

Once a preliminary set of SPICE parameters have been derived via the measurements outlined above, the draft model is verified in various simulation circuits that reproduce the measurement conditions, and adjustments made to appropriate parameters as required.

Model limitations

Please refer to Appendix A, which is a printout of the introductory text file on the Zetex SPICE model disc, and includes some comments on limitations.

In addition:

1. An important effect to consider for medium to high voltage device models is the **"Quasi-saturation Effect"** (Appendix B, reference 3).This effect is apparent as a two stage slope in the transistor's saturation region, and an abrupt fall in h_{FE} at medium to high collector currents. This feature has been addressed within later versions of PSPICE, which includes additional parameters to model this region of operation. Zetex SPICE models are intended to be of general application, so these additional parameters are not included within the SPICE models. Some experimentation is possible, using piecewise-linear (PWL) techniques with current sources within SPICE subcircuits, to enable different values of series resistance.

NOTE: It is recommended that simulations with high voltage models operating near to, or at their maximum operating current are thoroughly validated.

2. Device package. The SPICE fundamental device models are not aware of the package used to encapsulate the product, so resistance (other than that measured and accounted for within RC, RE etc), inductance and stray capacitance will not be included within the model. If these components are important within a particular application, then it is possible to create a subcircuit within SPICE that could incorporate the device model with additional passive components to represent parasitics.
Eg. Inductance for RF transistors and variable capacitance diodes, resistance for ESR components of RF diodes, and leakage resistance for MOSFETs. (Please refer to Figure 5 for an example of a

```
*    ZETEX   ZC830A   Spice Model          Last revision 4/3/92
*
.MODEL ZC830A D IS= 5.355E-15 N= 1.08 RS= 0.1161 XTI =3
+ EG= 1.11 CJO= 19.15E-12 M= 0.9001 VJ= 2.164 FC= 0.5
+ BV= 45.1 IBV= 51.74E-3 TT= 129.8E-9
* + ISR= 1.043E-12 NR= 2.01 (INCLUDE FOR LATER SPICE VERSIONS)
*
*NOTES: FOR RF OPERATION ADD PACKAGE INDUCTANCE 0F 2.5E-9H AND SET
*RS=0.68 FOR 2V, 0.60 FOR 5V, 0.52 FOR 10V OR 0.46 FOR 20V BIAS.
*
```

Figure 5
Zetex ZC830A Variable Capacitance Diode SPICE Model, Illustrating Parasitic and ESR Components.

variable capacitance diode with options provided for ESR and package inductance).

It also follows from the above that for dc or low frequency operation, and for those cases where a device model is not available for the package of interest, it may be worth considering models for electrically similar parts in other packages.

3. Breakdown voltage. The SPICE bipolar transistor models do not possess a breakdown voltage for any combination of terminals. This feature can be added within a subcircuit as above, by including a diode across the relevant terminals, with BV set to the minimum breakdown voltage specified on the transistor's datasheet. However, this will only indicate that there is current flow under overvoltage or transient conditions - it will **not** of course model the effects this would have on a device. This includes effects such as h_{FE} degradation due to reverse emitter-base current, heating and secondary breakdown effects, catastrophic breakdown of the gate oxide of MOSFETs, oscillation, and noise generation (at low values of avalanche current).

4. Safe Operating Area. Partially covered by the comments above on breakdown voltage, safe operating area (SOA) is not addressed by the SPICE model. So the designer would need to consider the usual SOA issues such as: maximum pulse currents (in reality limited by device gain h_{FE}/g_{FS} and the bond wire fusing current); thermal resistance of the basic silicon/package combination; and secondary breakdown loci. This can be done with reference to the models' circuit behaviour and the datasheet charts, but for final design validation full breadboard analysis is recommended.

5. Parametric variation. The models do not include Monte Carlo parameters so statistical analysis of circuit performance due to lot and device variation is not available.

NOTE: The model parameters are derived using nominal or mid-band value components.

Application Examples

This section presents several basic simulation circuits to demonstrate some of the functions that can evaluated using Zetex SPICE models. These circuits represent some of the more basic, typical applications that can benefit from using Zetex components, and may serve as a basis for further experimentation.

It is recommended that any design is thoroughly checked in hardware before committing the design to production, to avoid falling prey to any software generated optimism, or effects due to ***model*** limitations of the Zetex parts and other circuit components.

Discrete Component "Operational Amplifier". This form of circuit is very popular within the audio industry for low noise pre-amplification of mV level signals, particularly from low impedance sources such as moving coil transducers. The advantage of a discrete implementation is that it allows the designer to employ relatively large area, high gain input devices. These devices

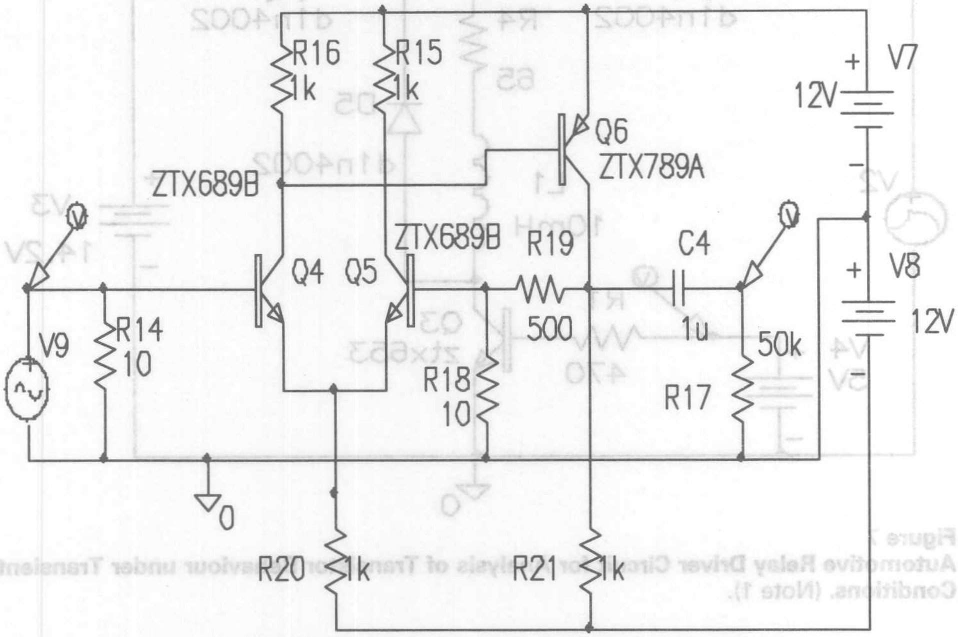


Figure 6
Discrete Component "Op-amp" using Matrix Geometry Transistors for Minimal Thermal Noise Generation. (Note 1).

present a low value of base spreading resistance to the input circuit, thereby minimising the amount of thermal noise generated. (Figure 6).

Automotive Relay Driver. Discrete transistors and Darlingtons have widespread appeal as a cost effective relay or solenoid driver. A first pass analysis of such a circuit may suggest that the operating conditions experienced by the device, (that is the level of performance demanded by the load under *nominal conditions*) are fairly benign. However, when transient events (such as generated by the many inductive loads sharing a common supply line) and environmental factors are considered, it is evident that the transistor driver deserves some attention.

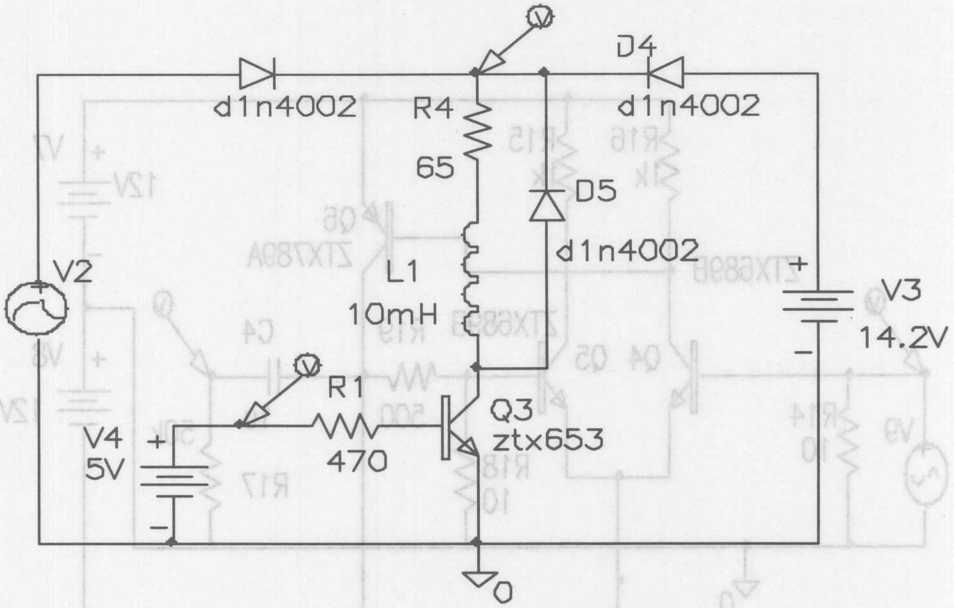


Figure 7
Automotive Relay Driver Circuit for Analysis of Transistor Behaviour under Transient Conditions. (Note 1).

As an example, the circuit shown in Figure 7 models the effect of a positive line transient on the operation of a simple relay driver while in the on-state. The resultant SPICE derived traces are given in Figure 8, which shows the collector current and collector-emitter voltage conditions imposed on the transistor. It can be seen that for the given bias condition, the transistor is unable to remain in the saturated state during the first few milli-seconds of the transient, but as the transient induced current falls, the transistor has sufficient h_{FE} to turn-on to a low $V_{CE(sat)}$. The period in which the device is within the linear operation region must be carefully considered, with respect to thermal and secondary breakdown compliance. SPICE could further be used to verify the transient power generated, and to model the effects of different base drives.

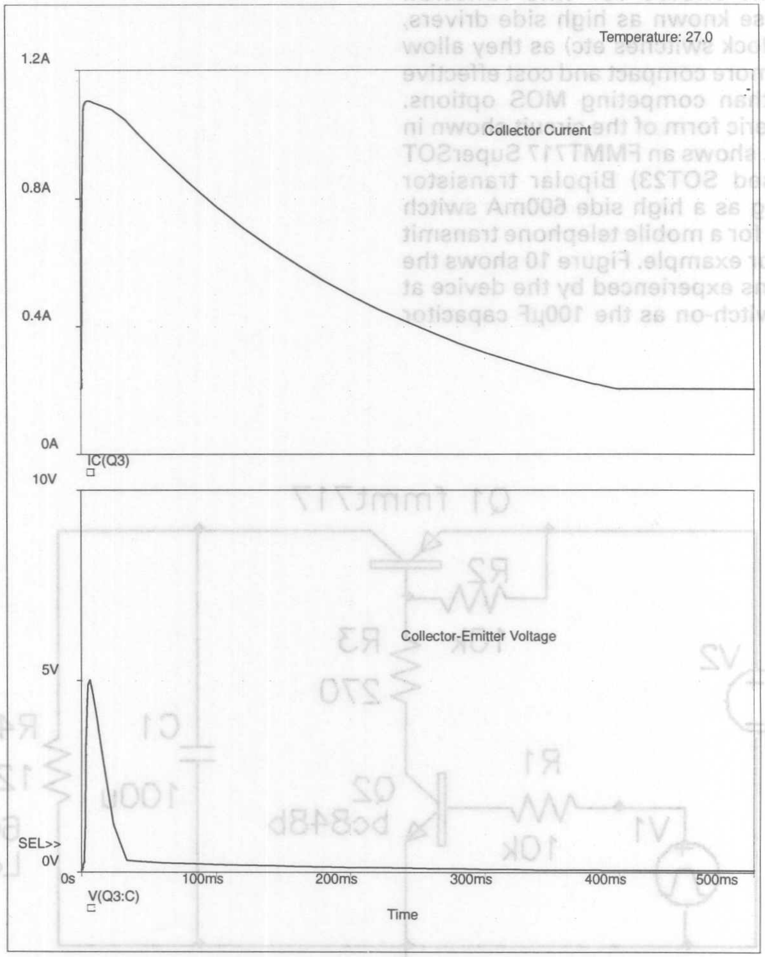


Figure 8
Automotive Relay Driver Circuit - SPICE Prediction of I_C and V_{CE} when subject to a + 80V, 230ms "Load Dump" Transient (Device "on" prior to Transient). (Note 1).

Positive Line Switch. High gain low $V_{CE(sat)}$ Bipolar transistors are an excellent choice for this function (otherwise known as high side drivers, circuit block switches etc) as they allow a much more compact and cost effective design than competing MOS options. The generic form of the circuit shown in Figure 9, shows an FMMT717 SuperSOT (optimised SOT23) Bipolar transistor operating as a high side 600mA switch -as used for a mobile telephone transmit switch for example. Figure 10 shows the conditions experienced by the device at initial switch-on as the 100µF capacitor charges.

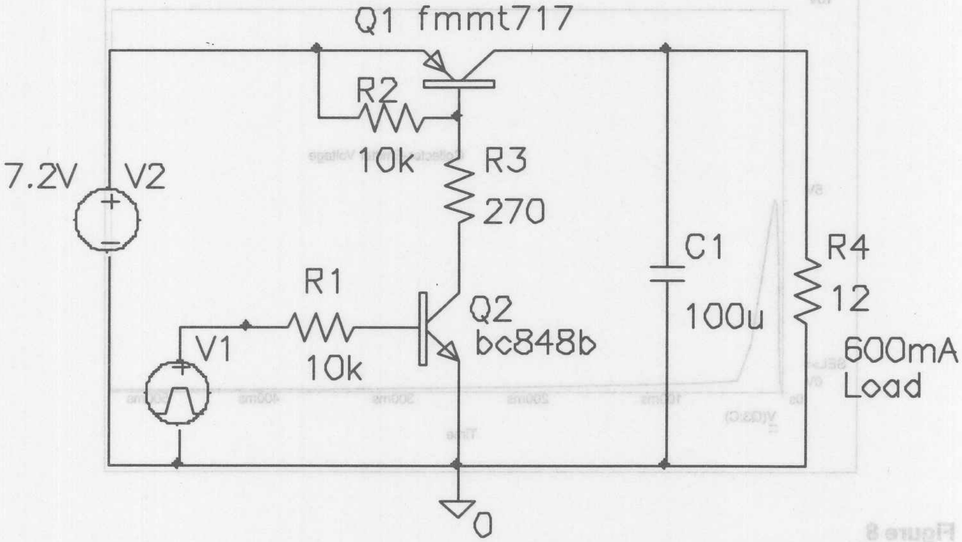


Figure 9
Positive Line Switch using Zetex FMMT717 SuperSOT SOT23 PNP Bipolar Transistor, as a 600mA Load Supply Switch. (Note 1).

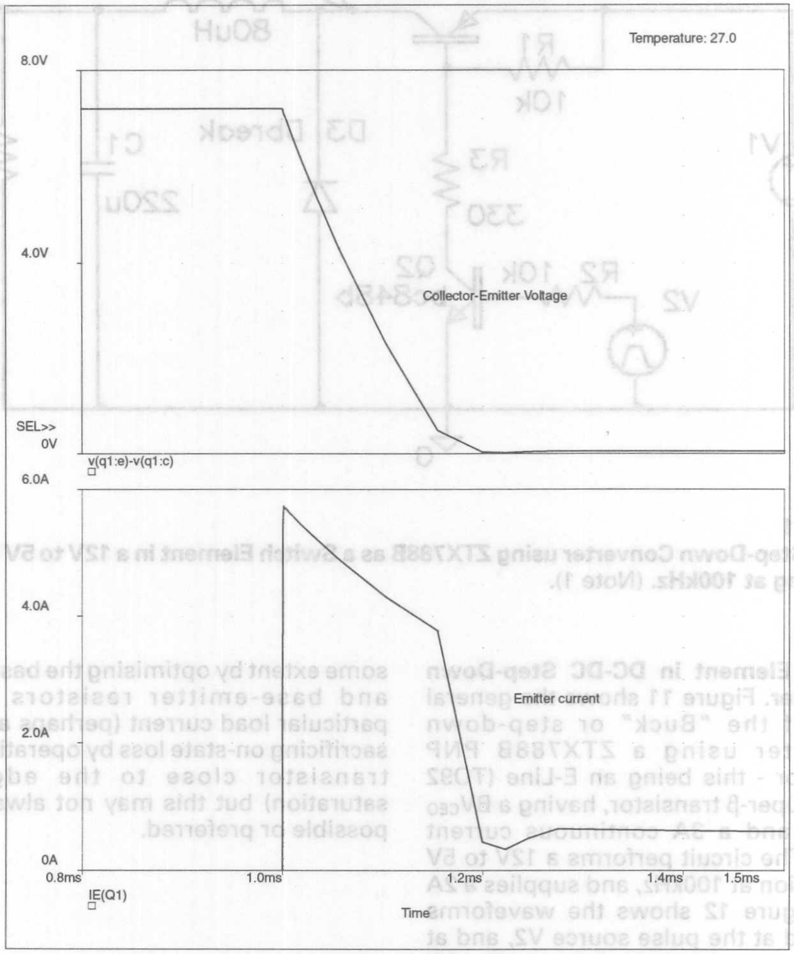


Figure 10
Positive Line Switch - SPICE Prediction of IC and VCE at Turn-on, showing Capacitive Charging Current in C1. (Note 1).

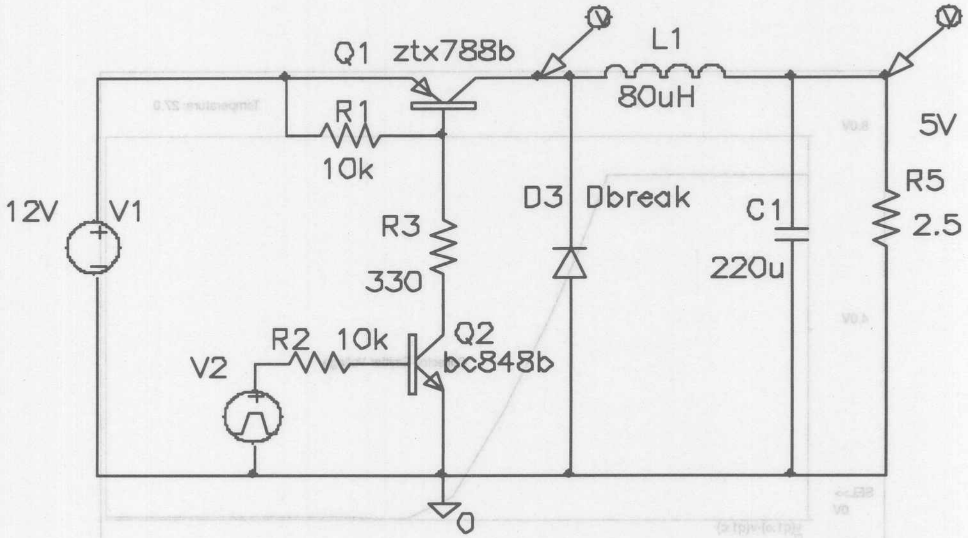


Figure 11
DC-DC Step-Down Converter using ZTX788B as a Switch Element in a 12V to 5V Circuit Operating at 100kHz. (Note 1).

Switch Element in DC-DC Step-Down Converter. Figure 11 shows the general form of the "Buck" or step-down converter using a ZTX788B PNP transistor - this being an E-Line (TO92 style) Super-β transistor, having a BV_{CEO} of 15V and a 3A continuous current rating. The circuit performs a 12V to 5V conversion at 100kHz, and supplies a 2A load. Figure 12 shows the waveforms recorded at the pulse source V2, and at the collector of the transistor, both with respect to 0V. The effect of the bipolar transistor storage time (ts) can be seen as a delay between the V2 transition to 0V, and the fall of the collector voltage. This storage time can be reduced to some extent by optimising the base bias, and base-emitter resistors for a particular load current (perhaps also by sacrificing on-state loss by operating the transistor close to the edge of saturation) but this may not always be possible or preferred.

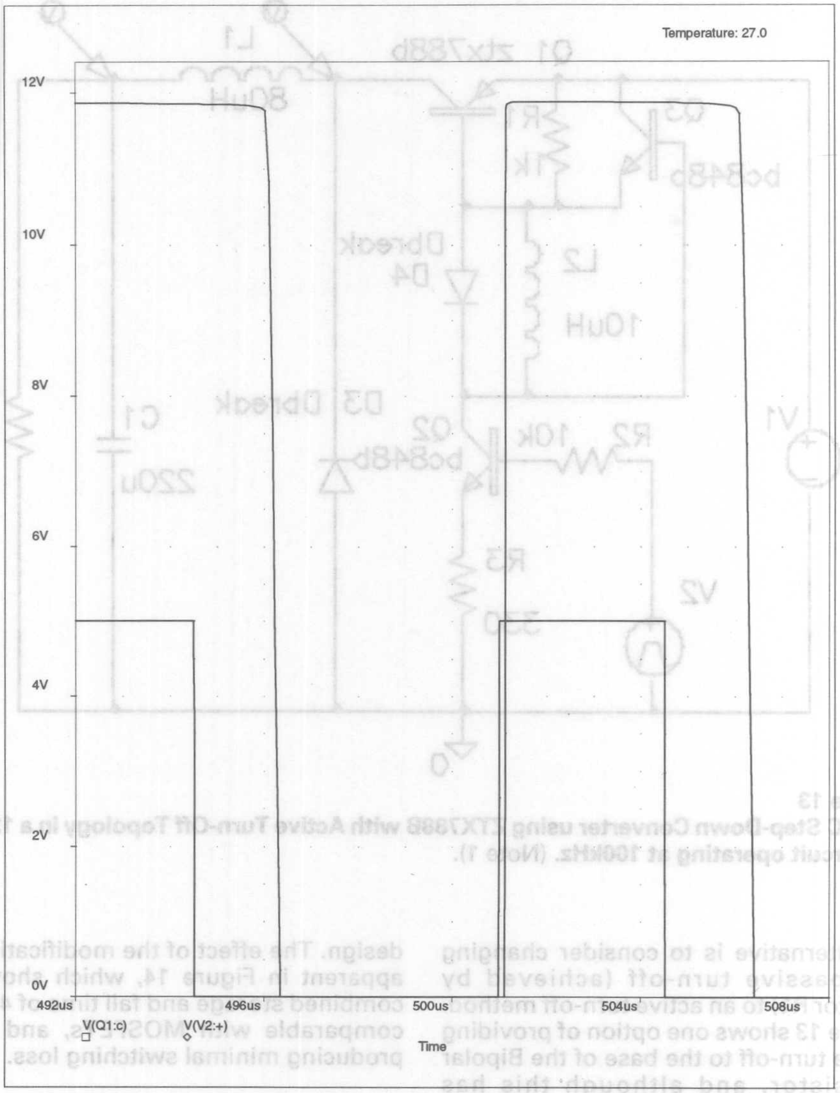


Figure 12
DC-DC Step-Down Converter - SPICE Prediction of ZTX788B Switching Waveforms. Traces show PWM drive (5V level) and Collector-to-0V Waveforms. Note Storage Time Effects. (Note 1).

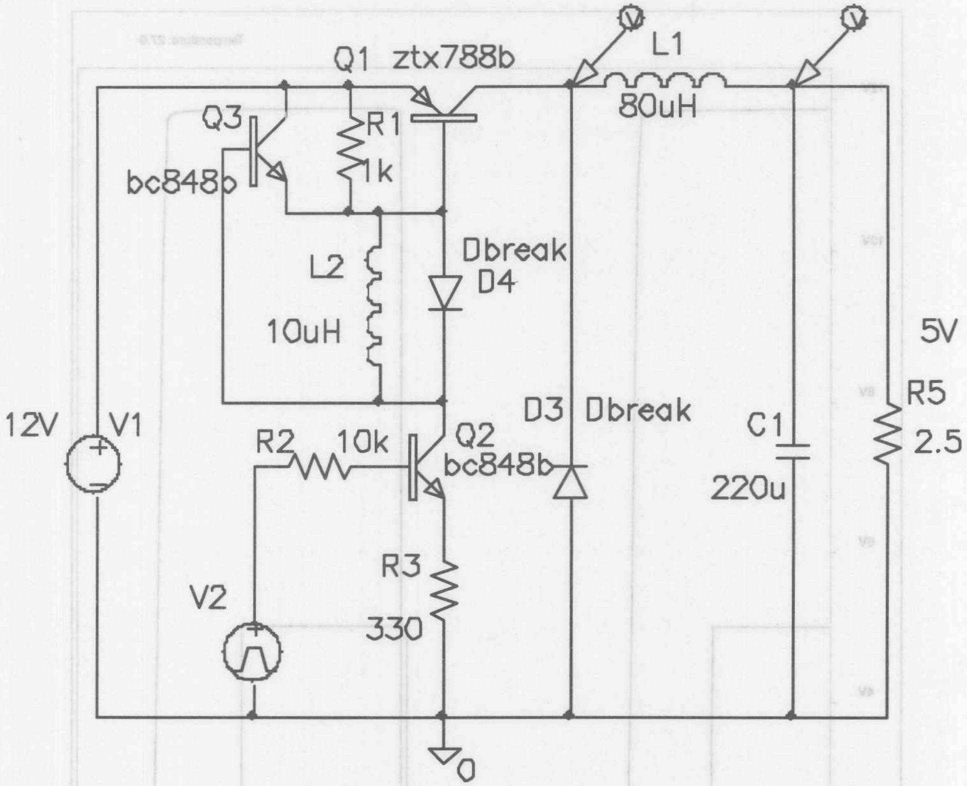


Figure 13
DC-DC Step-Down Converter using ZTX788B with Active Turn-Off Topology in a 12V to 5V Circuit operating at 100kHz. (Note 1).

An alternative is to consider changing the passive turn-off (achieved by resistor R1) to an active turn-off method. Figure 13 shows one option of providing active turn-off to the base of the Bipolar transistor, and although this has resulted in the addition of a few components, this implementation is more cost effective than would be the case with a TO220 MOSFET based design. The effect of the modification is apparent in Figure 14, which shows a combined storage and fall time of 40ns - comparable with MOSFETs, and also producing minimal switching loss.

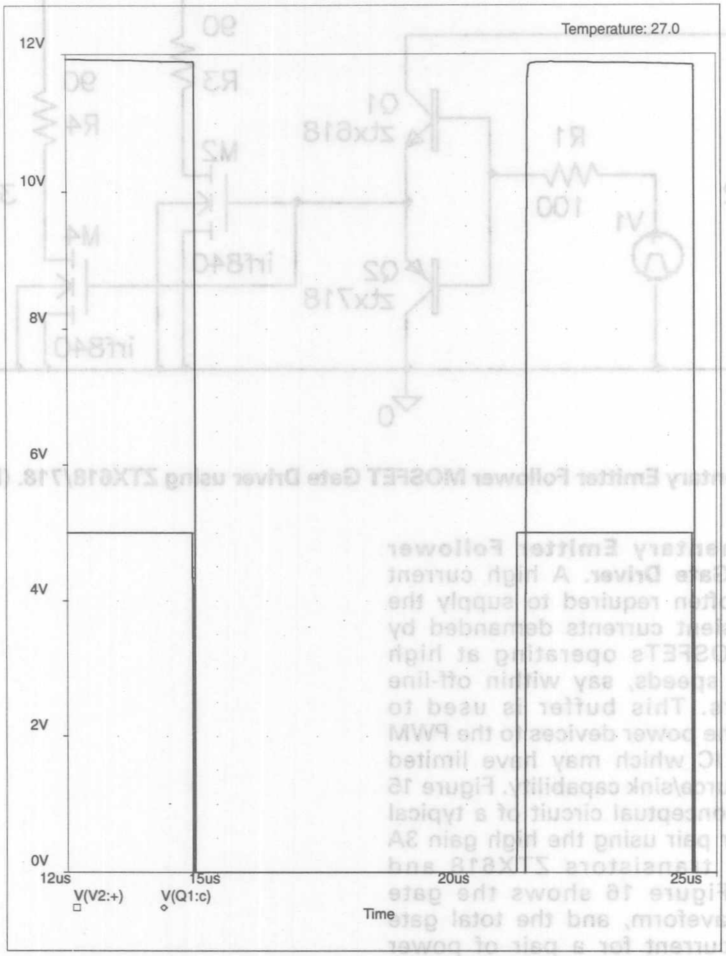


Figure 14
DC-DC Step-Down Converter - SPICE Prediction of ZTX788B Switching Waveforms. Traces show PWM Drive and Collector-to-0V Waveforms. Bipolar Transistor Turn-off Time = 50ns. (Note 1).

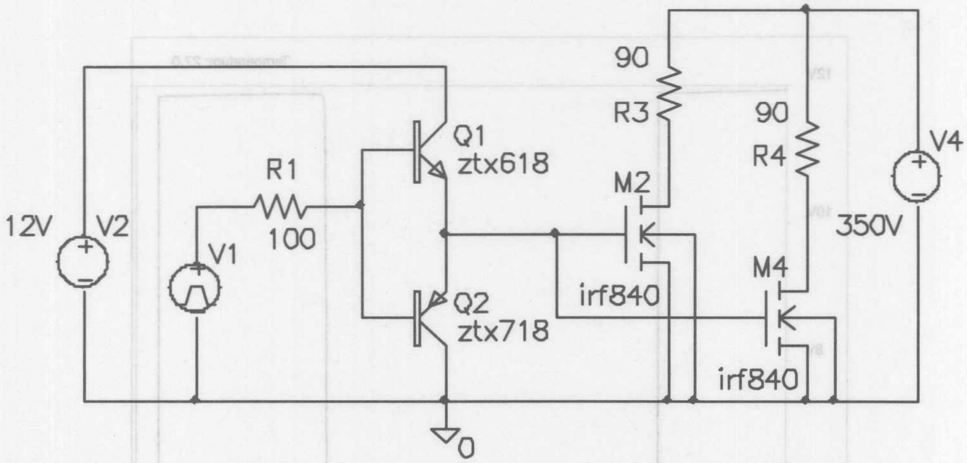


Figure 15
Complementary Emitter Follower MOSFET Gate Driver using ZTX618/718. (Note 1).

Complementary Emitter Follower MOSFET Gate Driver. A high current buffer is often required to supply the high transient currents demanded by Power MOSFETs operating at high switching speeds, say within off-line converters. This buffer is used to interface the power devices to the PWM controller IC which may have limited current source/sink capability. Figure 15 shows a conceptual circuit of a typical gate driver pair using the high gain 3A DC rated transistors ZTX618 and ZTX718. Figure 16 shows the gate voltage waveform, and the total gate charging current for a pair of power MOSFETs.

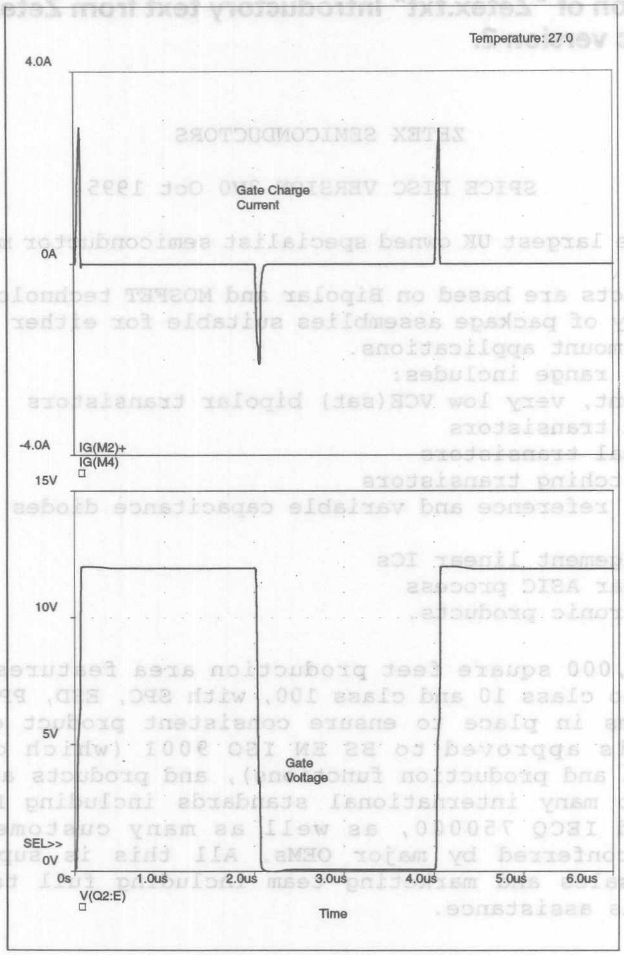


Figure 16
Complementary Emitter Follower MOSFET Gate Driver - SPICE Prediction of Gate Voltage and Current Waveforms. Gate Turn-On = 50ns. (Note 1).

Appendix A
Reproduction of "Zetex.txt" Introductory text from Zetex SPICE models disc version 2.

ZETEX SEMICONDUCTORS

SPICE DISC VERSION 2V0 Oct 1995

Zetex is the largest UK owned specialist semiconductor manufacturer.

Zetex products are based on Bipolar and MOSFET technologies offered in a variety of package assemblies suitable for either through-hole or surface mount applications.
The product range includes:
- High current, very low VCE(sat) bipolar transistors
- Darlington transistors
- Small signal transistors
- RF and switching transistors
- Switching, reference and variable capacitance diodes
- MOSFETs
- Power management linear ICs
- A 20V linear ASIC process
- Opto-electronic products.

Zetex's 60,000 square feet production area features cleanrooms operating to class 10 and class 100, with SPC, ESD, PPM, FMEA, and FIT programs in place to ensure consistent product quality. The facility is approved to BS EN ISO 9001 (which covers both development and production functions), and products are routinely supplied to many international standards including BS9300, CECC 50000, and IECQ 750000, as well as many customer specific approvals conferred by major OEMs. All this is supported by a dedicated sales and marketing team including full technical and applications assistance.

For more information on the company and its products, please contact your nearest sales office.

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Zetex Spice Models

Introduction

Welcome to version 2 of the Zetex disc of Spice device models. This release contains Spice models for 108 Zetex device parents that source approximately 280 device types. These types include switching and varicap diodes, bipolar (high current low VCE(sat)),small signal bipolar, RF bipolar, bipolar Darlington and MOSFET transistors. This range is continuously under expansion as new products are introduced and retrospective models are generated for existing products.

This latest version of the Spice disc also includes a symbol library that allows the models to be used with the Windows versions of PSpice. Further information on the symbol library,including installation instructions, will be found in an additional text file in the root directory of this disc. This file is called ZETEXSYM.TXT.

The Spice models supplied on this disc are accessible in three ways:

1) The directory ZTXMODS contains a separate Spice model file for each Zetex device type for which a model is presently available. New model releases between major revisions of the Spice disc will be stored in the NEWMODS directory.

2) The directory ZTXLIBS contains the file ZMODELS.LIB in which all available Zetex device models are collected into a single file. New model releases between major revisions of the Spice disc will be stored within the ZNEWMODS.LIB library.

3) The directory ZTXWIN contains a symbol library (ZETEXM.SLB) that enables Windows versions of PSpice to use the Zetex SPICE models.

ZTXMODS and NEWMODS Model Files

Each of these files is a Spice model for a single Zetex device. They can be loaded into your simulation simply by employing the Spice command <.INCLUDE Device_name.MOD>. Only the device types specifically required by the circuit under simulation need be included in this way. All diode models and all but one of the bipolar transistor models are simple <.MODEL> files. However, the model for the FMMT597Q, all Darlington transistors and all MOSFET models are multi-component subcircuits and so are supplied as <.SUBCKT> files.

The diode models should be included in circuit files using the normal Spice reference: <Dnum Anode_node Cathode_node Device_name>.

Bipolar transistor models should be included using <Qnum Collector_node Base_node Emitter_node Device_name>.

All other models should be referenced as subcircuits,ie in the form <Xnum Collector_node Base_node Emitter_node Device_name> for the FMMT597Q and all Darlington transistors,and as <Xnum Drain_node Gate_node Source_node Device_name> for all MOSFET models.

The ZMODELS.LIB and ZNEWMODS.LIB Library Files

To save disc and directory space,some users may prefer to use the model libraries. For later Spice versions the ZMODELS.LIB and ZNEWMODS.LIB libraries are available. These are simply collections of all Zetex Spice models exactly as they appear in the individual model directories. By using the statement <.LIB ZMODELS.LIB> and <.LIB ZNEWMODS.LIB>, Spice will be able to access any model within the libraries without the need for multiple <.INCLUDE> statements.

Note that all subcircuits, be they in the library files or the individual model files use the same connection sequence as Spice uses for single element models, thus easing their use.

Model Parameters and Limitations

Bipolar Models

All bipolar transistor and Darlington models are based on the Spice modified Gummel-Poon model. Following is a typical model for a single transistor:-

```
ZETEX ZTX109 Spice model Last revision 4/90
*
.MODEL ZTX109 NPN IS=1.8E-14 ISE=5.0E-14 NF=.9955 BF=400 BR=35.5
+IKF=.14 IKR=.03 ISC=1.72E-13 NC=1.27 NR=1.005 RB=.56 RE=.6 RC=.25
+VAF=80 VAR=12.5 CJE=13E-12 TF=.64E-9 CJC=4E-12 TR=50.72E-9
MJC=.33
*
```

A brief guide to the effect of each model element :-

IS and NE controls Icbo and where hFE falls with high Ic.
ISE and NE control the fall in hFE that occurs at low Ic.
BF controls peak forward hFE.
BR controls peak reverse hFE ie collector and emitter reversed.
IKF controls where hFE falls at high collector currents.
IKR controls where reverse hFE falls at high emitter currents.
ISC and NC controls the fall of reverse hFE at low currents.
RC, RB and RE add series resistance to these device terminals.
VAF controls the variation of collector current with voltage when the transistor is operated in its linear region.
VAR the reverse version of VAF.
CJC, VJC and MJC control Ccb and how it varies with Vcb.
CJE controls Cbe.
TF controls Ft and switching speeds.
TR controls switching storage times.

The standard bipolar transistor Spice model includes a parameter that allows BF, the hFE parameter, to vary with temperature. This parameter is called XTB and defaults to zero, Eg. no temperature dependence. If hFE temperature effects are of interest, then the following values may be used to provide an estimate, or a starting point for further investigation. It is suggested that the appropriate data sheet hFE profile is examined, and a Spice test circuit created that simulates the device in question and generates a set of hFE curves. Two or three such iterations should normally be sufficient to define a value for XTB in each case.

Polarity	XTB
NPN	1.6
PNP	1.9

Please remember that these notes are only a rough guide as to the effect of model parameters. Also, many of the parameters are interdependent so adjusting one parameter can affect many device characteristics.

At Zetex, we have endeavoured to make the models perform as closely to actual samples as possible but some compromises are forced which can result in simulation errors under some circumstances.The main areas of error observed so far have been:-

1. Spice is often over optimistic in the hFE a transistor will give when operated above it's data sheet current ratings. This is particularly true for a high voltage transistor operated at a low collector-emitter voltage.

2. Spice can be pessimistic when predicting switching storage time when current is extracted from the base of a transistor to speed turn-off.

Darlington Models

These are subcircuits using a standard transistor model. A Darlington model looks like:-

```
*ZETEX BCX38B Darlington Spice Subcircuit Last revision 4/9/90
*
.SUBCKT BCX38B 1 2 3
*              C B E
Q1 1 2 4 SUB38B
```

```
Q2 1 4 3 SUB38B 12.75
*

.MODEL SUB38B NPN IS=1.1E-14 ISE= etc
+ etc
.ENDS BCX38B
```

Note that because Zetex Darlingtons are monolithic, the two transistors used are identical in all respects other than size. (The number at the end of the Q2 line multiplies the size of the SUB38B transistor by 12.75 - the ratio of the areas of the input and output transistors for this device).

MOSFET Models

None of Spice's standard MOSFET models fit the characteristics of vertical MOSFETs too well. Consequently the models of Zetex MOSFETs supplied on this disc have been made using subcircuits that include additional components to improve simulation accuracy.

A typical MOSFET model:-

```
*ZETEX ZVN4106 MOSFET Spice Subcircuit Last revision 11/91
*
.SUBCKT ZVN4106  3 4 5
*Nodes           D G S
M1 3 2 5 5 MOD1 L=1 W=1
RG 4 2 343
RL 3 5 6E6
D1 5 3 DIODE1
.MODEL MOD1 NMOS VTO=2.474 RS=1.68 RD=0.0 IS=1E-15 KP=-.296
+CGSO=23.5P CGDO=4.5P CDB=35.5P PB=1 LAMBDA=267E-6
.MODEL DIODE1 D IS=1.254E-13 N=1.0207 RS=0.222
.ENDS ZVN4106
*
```

In the NMOS model,

VTO defines Vgs(th).
RS and RD add series terminal resistance.
IS controls the behaviour of the model's body diode.
KP controls Gm
LAMBDA controls variation of drain current with drain-source

voltage when operated in the linear region.
CGDO controls Crss.
CGSO controls Ciss.
CBD controls Coss.

Added to Spice's standard NMOS model are a gate resistor to control switching speeds, a drain-source resistor to control leakage and a drain-source diode to accurately reflect the performance of the MOSFET's body diode.
The MOSFET models mirror the performance of the real devices well in most areas. One area not covered however is the way that Crss and Coss varies with drain-source voltage. Thus if the models are used at a drain-source voltage well away from data sheet capacitance definition voltages, and capacitance is critical, then the values used for CGSO and CGDO may need adjustment.

Diode Models

Diodes from Zetex's Switching and Varicap range are presently modelled on this disc. They use a standard Spice diode model and a typical file appears as follows:-

```
*ZETEX ZC830A Spice Model Last revision 4/3/92
*
.MODEL ZC830A D IS=5.355E-15 N=1.08 RS=0.1161 XTI=3
+ EG=1.11 CJO=19.15E-12 M=0.9001 VJ=2.164 FC=0.5
+ BV=45.1 IBV=51.74E-3 TT=129.8E-9
* + ISR=1.043E-12 NR=2.01 (INCLUDE FOR LATER SPICE VERSIONS)
*
*NOTES: FOR RF OPERATION ADD PACKAGE INDUCTANCE 0F 2.5E-9H AND SET
*RS=0.68 FOR 2V, 0.60 FOR 5V, 0.52 FOR 10V OR 0.46 FOR 20V BIAS.
*
```

In this model,

IS contols forward and reverse current against voltage.
N controls forward current against voltage.
RS controls forward voltage at high current.
CJO, M and VJ control variation of capacitance with voltage.
BV and IBV control reverse breakdown characteristics.
TT controls switching reverse recovery characteristics.

ISR and NR (if activated) control reverse biased leakage.

For operation at RF (which would be the norm for a varicap diode) it is recommended that a 2.5nH series inductor be added as an extra circuit element to correct for the inherent package inductance. Also, to give the varicap diode it's correct Q, the model value for RS should be changed to the appropriate figure suggested in the NOTES for the expected reverse bias.

The switching diode models may include a constant value capacitor rather than the parameters CJO etc if chip and packaging strays greatly exceed true junction capacitance values.

Further Information

Zetex's library of Spice models is being continuously updated so if the model you require does not appear on this disc, please contact us at the address given below or through your local Zetex office. If you have any problems with the models supplied here, you may use the same address to request applications assistance.

Zetex plc
Fields New Road, Chadderton, Oldham, OL9 8NP, United Kingdom
Telephone (44)161-627-5105 (Sales),
(44)161-627-4963 (General Enquires)
Facsimilie: (44)161-627-5467

Appendix B

References

1. "An Integral Charge Control Model of Bipolar Transistors", Gummel and H.C. Poon, Bell Syst. Tech. J., Vol. 49, May 1970.

2. "Modeling the Bipolar Transistor" CAD for Electronic Circuits , Volume 1, Ian E. Getreu. Elsevier Scientific Publishing Company. ISBN 0-444-41722-2.

3. "MicroSim PSPICE A/D Circuit Analysis Reference Manual", Analog Devices, Pages 170-178.

4."SPICE- A Guide to Circuit Simulation & Analysis Using PSPICE" , Paul W. Tuinenga. Prentice-Hall, Inc. ISBN 0-13-747270-6.

5. "Measuring Small Signal Common Emitter Current Gain of Transistors at High Frequencies" ASTM Standard, Vol. 10.04 Electronics, F632-79.

Note 1: Circuit diagrams and waveforms reproduced from PSpice "Schematics" and "Probe" screens, with permission of MicroSim Corporation.

ZETEX

Application Note 24
Issue 2 January 1996

An Introduction to the SM-8 Package

Mike Townson

Introduction

Over recent years the benefits for companies to move to surface mount technology has lead to significant growth in the component industry. The requirements on the component suppliers is to provide either smaller components or components containing more than one device. The aim being to reduce PCB size and the number of component placements, thus reducing overall costs. Zetex is expanding it's range of surface mount packages to meet these demands. The first in this new series of packages to be introduced is the SM-8.

What is SM-8?

The SM-8 package has evolved from the industry standard SOT223 surface mount package. The development of a new lead frame, whilst maintaining the physical outline of the standard package, effectively gives two transistors in a package initially designed for one. Figure 1 illustrates the package.

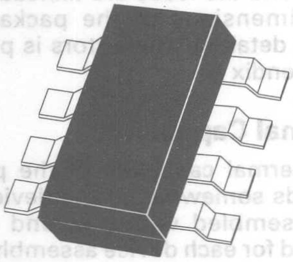


Figure 1
The SM-8 package.

Package Construction

The SM-8 is an eight pin device that can be configured in two ways due to different lead frame options. Both frame options enable two totally independent devices to be assembled into the same package.

The first frame option has eight independent pins enabling two devices that require four connections to be assembled in the package, for example two high side drivers, such as the Zetex ZHD100 BiMos switch.

The second frame option has two sets of two pins connected internally, effectively allowing the assembly of two devices requiring three connections, for example two Bipolar transistors or MOSFETs. The two common pins are internally connected to the frame onto which the transistor die is attached, and through which the collector or drain connection is made. This provides a low thermal resistance, and therefore allows a good transfer of heat away from the semiconductor chip through the frame, and onto the board or substrate.

The body of the device is a moulded epoxy and the leads are tin/lead plated. The dimensions of the package and pinout detail for transistors is provided in Appendix A.

Thermal Capability

The thermal capability of the package depends somewhat on the devices that are assembled within it, and will be detailed for each device assembly on the appropriate datasheet. However for illustrative purposes, the dual Bipolar transistor ZDT1048 device is used here as an example. Figures 2 and 3 show the transient and DC thermal resistance response for the ZDT1048 device, when one and two devices are powered respectively, and show DC values of thermal resistance of 55.6 and 45.5°C/W. This leads to a package power handling capability of 2.75W at an ambient temperature of 25°C, when both devices are turned on equally. If the circuit operation is such that only one of the devices is on at once, then the capable dissipation is 2.25W. The data for the above was derived from thermal resistance measurements with the package mounted on a standard FR4 PCB with a copper area of 2 inches square. As with any surface mount component, the actual thermal resistance achieved depends on many factors including the board area, the board material, proximity of passive and other components, and whether the board is single, double sided or multi-layer.

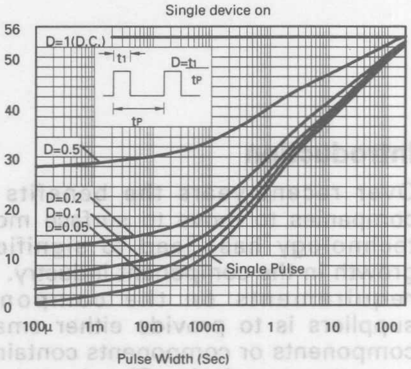


Figure 2
Transient Thermal Resistance Curves for the ZDT1048 - Single Device on.

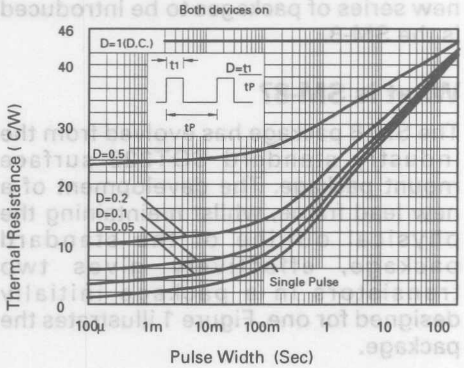


Figure 3
Transient Thermal Resistance Curves for the ZDT1048 - Both Devices on.

The latter is particularly effective in dissipating heat as the internal copper traces encourage lateral heat flow within the board, therefore possibly increasing the area from which the heat is dissipated. Figure 4 shows how the thermal resistance varies with single copper sided FR4 PCB area.

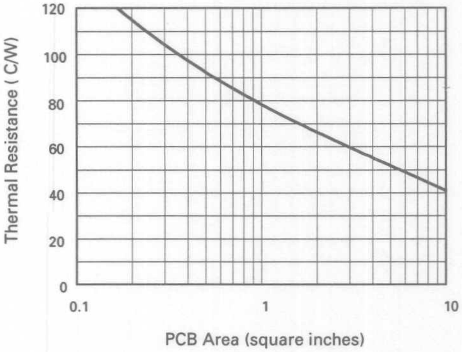


Figure 4
Thermal Resistance vs PCB Area (Single Copper Layer on FR4).

Electrical Specification

Theoretically any combination of two chips from the Zetex range of components can be assembled in the SM-8 package, however, commercial implications will be taken into consideration before introduction. This gives a potential current handling capability up to 5 Amps continuous with 20 Amps pulsed and voltages ranging from 10 to 450 Volts. Appendix B reproduces a datasheet for one of the dual transistor products - the ZDT1048. This device has been developed specifically for service within LCD Backlight Inverters.

Product Range

The product range is still in its infancy. New products are being introduced on a gradual basis as new opportunities are identified. An introductory range has been generated to demonstrate the options that are available to design engineers, This includes dual NPN or PNP transistors, NPN and PNP combinations, MOSFET combinations, dual high side drivers, and various Linear IC combinations.

Future developments will include dual IGBTs, and linear ICs/discrete component combinations.

Overall the SM-8 package offers Zetex and it's customers the flexibility to provide innovative and cost effective circuit solutions.

Applications

For discrete component combinations the opportunities will lie in designs using push pull circuits, half or full bridges, Royer converters or high side drivers. These can be found in such applications as compact fluorescent ballasts, emergency lighting, LCD backlighting, motor drives and siren drivers.

As an example of how the SM-8 package can transform a product, Figure 5 shows how Mitel Semiconductors have been able to reduce the size of their SLIC hybrid (part # MH88615) by replacing 4 SOT223 packaged devices with 2 SM-8s. The SLIC (Subscriber Line Interface Card) provides a complete interface between a switching system and a subscriber loop.

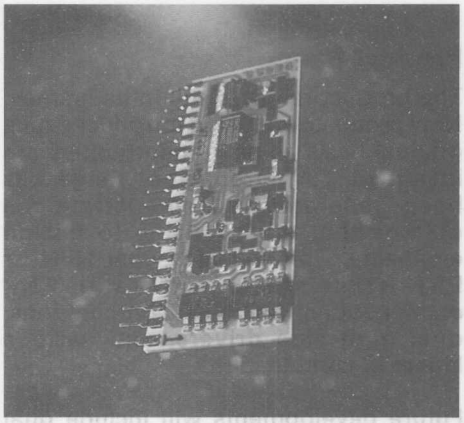
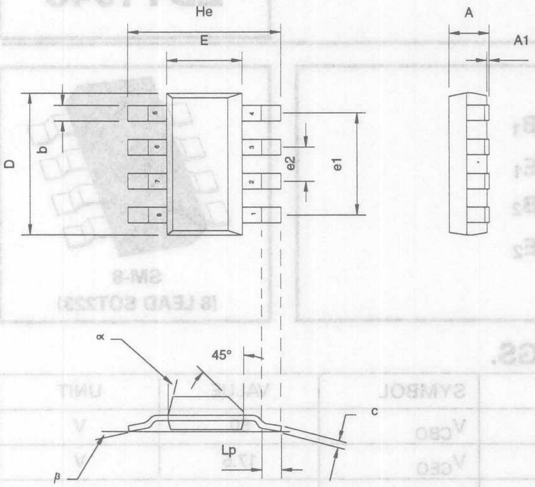


Figure 5
Mitel SLIC Hybrid using SM-8 Dual transistors.

Future Developments

Evaluation is ongoing of suitable four chip assemblies including a 'H'-Bridge configuration (4 transistors), an anti-parallel diode protected half-bridge (2 transistors and 2 diodes) and Schottky diode bridges. Plans are already underway to develop an 8 lead version of the smaller SOT23 package, again with the aim of offering two independent devices in one package. This part will be called the SSM-8, and it is expected that using variants of the SuperSOT geometry will enable the package to house two 15V devices, individually capable of conducting up to 3A continuous, and 12A under pulsed conditions.

Appendix A
Dimensional and Pinout details



PIN NUMBER	BIPOLARS	MOSFETS
1	E2	S2
2	B2	G2
3	E1	S1
4	B1	G1
5	C1	D1
6	C1	D1
7	C2	D2
8	C2	D2

Pinout details for dual Bipolar and MOSFET products.

Dim	Millimetres			Inches		
	Min	Typ	Max	Min	Typ	Max
A	–	–	1.7	–	–	0.067
A1	0.02	–	0.1	0.0008	–	0.004
b	–	0.7	–	–	0.028	–
c	0.24	–	0.32	0.009	–	0.013
D	6.3	–	6.7	0.248	–	0.264
E	3.3	–	3.7	0.130	–	0.145
e1	–	4.59	–	–	0.180	–
e2	–	1.53	–	–	0.060	–
He	6.7	–	7.3	0.264	–	0.287
Lp	0.9	–	–	0.035		
α	–	–	15°	–	–	15°
β	–	10°	–	–	10°	–

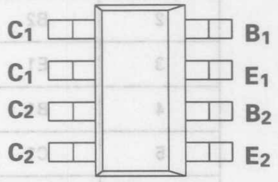
Dimensional detail.

Appendix B
Sample Datasheet

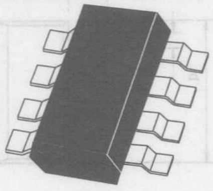
SM-8 DUAL NPN MEDIUM POWER HIGH GAIN TRANSISTORS

ISSUE 2 - FEBRUARY 1996

ZDT1048



PARTMARKING DETAIL – T1048



SM-8
(8 LEAD SOT223)

ABSOLUTE MAXIMUM RATINGS.

PARAMETER	SYMBOL	VALUE	UNIT
Collector-Base Voltage	V_{CBO}	50	V
Collector-Emitter Voltage	V_{CEO}	17.5	V
Emitter-Base Voltage	V_{EBO}	5	V
Peak Pulse Current	I_{CM}	20	A
Continuous Collector Current	I_C	5	A
Base Current	I_B	500	mA
Operating and Storage Temperature Range	T_j:T_{stg}	-55 to +150	°C

THERMAL CHARACTERISTICS

PARAMETER	SYMBOL	VALUE	UNIT
Total Power Dissipation at T_{amb} = 25°C* Any single die "on" Both die "on" equally	P_{tot}	2.25 2.75	W W
Derate above 25°C* Any single die "on" Both die "on" equally		18 22	mW/ °C mW/ °C
Thermal Resistance - Junction to Ambient* Any single die "on" Both die "on" equally		55.6 45.5	°C/ W °C/ W

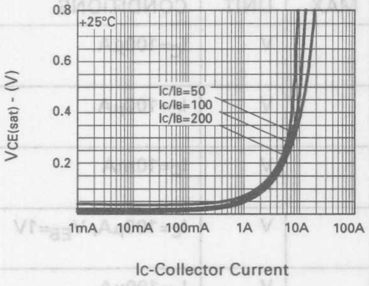
* The power which can be dissipated assuming the device is mounted in a typical manner on a PCB with copper equal to 2 inches square.

ELECTRICAL CHARACTERISTICS (at Tamb = 25°C unless otherwise stated).

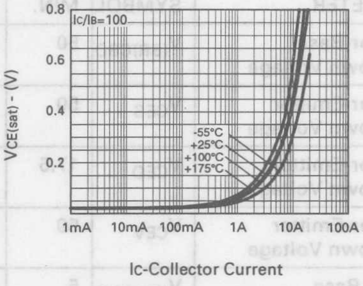
PARAMETER	SYMBOL	MIN.	TYP.	MAX.	UNIT	CONDITIONS.
Collector-Base Breakdown Voltage	$V_{(BR)CBO}$	50	85		V	I_C=100µA
Collector-Emitter Breakdown Voltage	V_{CES}	50	85		V	I_C=100µA
Collector-Emitter Breakdown Voltage	V_{CEO}	17.5	24		V	I_C=10mA
Collector-Emitter Breakdown Voltage	V_{CEV}	50	85		V	I_C=100µA, V_{EB}=1V
Emitter-Base Breakdown Voltage	$V_{(BR)EBO}$	5	8.7		V	I_E=100µA
Collector Cutoff Current	I_{CBO}		0.3	10	nA	V_{CB}=35V
Emitter Cutoff Current	I_{EBO}		0.3	10	nA	V_{EB}=4V
Collector Emitter Cutoff Current	I_{CES}		0.3	10	nA	V_{CES}=35V
Collector-Emitter Saturation Voltage	$V_{CE(sat)}$		27 55 120 200 200	45 75 160 240 300	mV mV mV mV mV	I_C=0.5A, I_B=10mA* I_C=1A, I_B=10mA* I_C=2A, I_B=10mA* I_C=5A, I_B=100mA* I_C=5A, I_B=50mA*
Base-Emitter Saturation Voltage	$V_{BE(sat)}$		1000	1100	mV	I_C=5A, I_B=100mA*
Base-Emitter Turn-On Voltage	$V_{BE(on)}$		900	1000	mV	I_C=5A, V_{CE}=2V*
Static Forward Current Transfer Ratio	h_{FE}	280 300 300 250 50	440 450 450 300 80	 1200 		I_C=10mA, V_{CE}=2V* I_C=0.5A, V_{CE}=2V* I_C=1A, V_{CE}=2V* I_C=5A, V_{CE}=2V* I_C=20A, V_{CE}=2V*
Transition Frequency	f_T		150		MHz	I_C=50mA, V_{CE}=10V f=50MHz
Output Capacitance	C_{obo}		60	80	pF	V_{CB}=10V, f=1MHz
Switching Times	t_{on}		120		ns	I_C=4A, I_B=40mA, V_{CC}=10V
	t_{off}		250		ns	I_C=4A, I_B=±40mA, V_{CC}=10V

*Measured under pulsed conditions. Pulse width=300µs. Duty cycle ≤ 2%

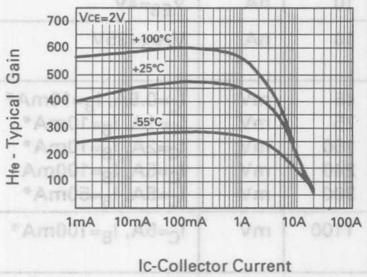
TYPICAL CHARACTERISTICS



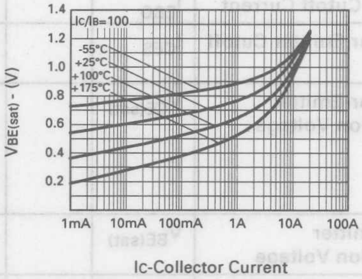
VCE(sat) v IC



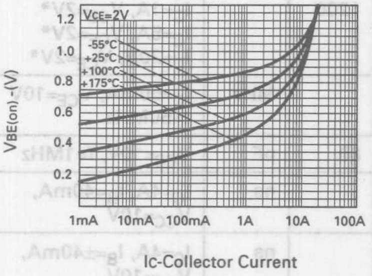
VCE(sat) v IC



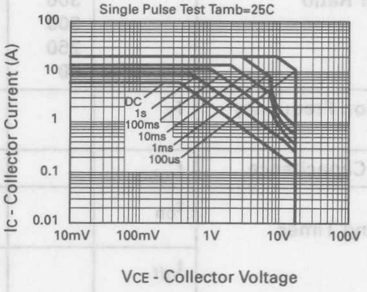
hFE v IC



VBE(sat) v Ic



VBE(on) v IC



Safe Operating Area

ZETEX

Application Note 25
Issue 2 January 1996

Soldering Surface Mount Devices

Whilst Surface Mount Devices (SMD) offer many important advantages, it should be recognised by potential users that there is a technological change required in moving from through-hole (leaded component) to surface mount assembly.

Unlike leaded components, which are mechanically supported by insertion of the leads through the board, SMDs rely on the solder joints alone to provide both the mechanical support and the electrical connection.

If the board is to be inverted for wave soldering then the devices are usually secured by an (epoxy) adhesive prior to the actual soldering operation. In the case of reflow soldering, where the devices remain uppermost, then they are usually held in place by solder paste which has been screen-printed onto the PCB prior to placement.

Processing: general

High Temperature Exposure:
SMDs, having shorter leads and being mounted on the side of the PCB exposed directly to the heat source, are at a much higher risk of damage from thermal shock. Care should therefore be taken to minimise both the thermal gradients and the maximum solder temperature. Low profile, high pin count, fine pitch devices are at greatest risk and their presence on the board will ultimately dictate the soldering parameters.

Soldering Conditions:
Not only will this depend on the component mix (as mentioned above), but also on board size and total thermal mass. For users who assemble a variety of different boards, it may be necessary to tailor the solder profiles for those boards which exhibit significant variations in the soldering characteristics. Fine pitch devices are normally restricted to reflow soldering processes where the solder is applied to the boards prior to device placement and then the whole board assembly is heated above the melting point of the solder.

Fluxes:
Highly active fluxes are not recommended and should not be required provided that the devices have not been exposed to any extreme storage conditions. Strongly acidic fluxes of the type associated with aqueous cleaning systems, should be treated with caution as they have been linked to device corrosion. Standard board cleaning techniques should not

present device problems provided that the boards are properly dried to remove solvents or water residues.

Reflow Soldering

1. Ramp-up:
During this stage the solvents are evaporated from the solder paste. Care should be taken to prevent rapid oxidation (or paste slump) and the subsequent possibility of solder bursts through aggressive solvent outgassing.

A typical heating rate is 2°C/sec.

2. Pre-heat / soak:
A constant temperature zone which performs the dual role of activating the flux whilst achieving a uniform board temperature. The objective is to ensure even wetting characteristics across all the devices on the board.

Typical soak conditions are: 30-150 secs at 150-170°C.

3. Reflow zone:
Maximum soldering temperatures will be determined by the most temperature sensitive devices on the PCB, which will typically be the large area, high pin count packages. If the temperature is too high then the devices may be damaged by mechanical stressing due to thermal mis-match or there may be problems due to excessive solder oxidation. Too low a temperature may mean that some of the devices fail to achieve the minimum soldering conditions. However, caution should be taken in extending the dwell time to compensate, as excessive time at temperature can enhance the formation of intermetallic compounds at the lead / board interface and may lead to early mechanical failure of the joint. High density circuits may require finer tuning of the profile because of their higher thermal mass.

Maximum soldering temperatures for most boards will be in the range 215-220°C, with a limit of 235°C.

4. Cooling:
Again, there is a compromise to be achieved: steep thermal gradients may give rise to excessive thermal shock, but fast cooling will promote a finer grain structure and a more crack resistant joint.

Figure 1 indicates the recommended soldering profile.

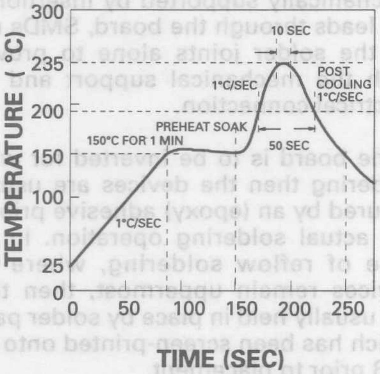


Figure 1
Reflow Soldering Profile.

Numerical values, along with suggested operating ranges are summarised in Table 1.

Heating rate during pre-heat	Typical 1-2, max 4°C / sec
Pre-heat temperature	Typical 150, max 170°C
Pre-heat dwell time	Typical 60, max 180 sec
Liquidus time above 200°C	30-90 sec

Table 1
Reflow Soldering Parameters.

Wave (Flow) Soldering

In this process it is normal for the SMDs to be secured to the board by an epoxy-based adhesive prior to soldering. Too much adhesive may encroach upon the pads or component leads, whereas too little may not be sufficient to prevent the component moving before the adhesive has cured.

The solder pad must extend beyond the end of the component (lead) and be large enough to pick up sufficient solder from the wave to make a satisfactory joint.

Care should be taken to prevent shadowing by other components and also to avoid bridging of leads. For this reason, wave soldering is not normally the preferred method for fine pitch devices. It is more usually employed as one of the two (or more) soldering operations which take place during the assembly of mixed technology boards.

Figure 2 indicates the recommended soldering profile.

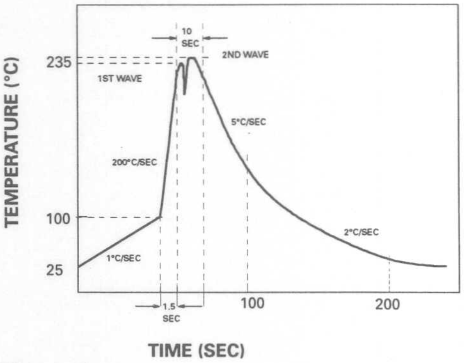


Figure 2
Wave Soldering Profile.

Numerical values, along with suggested operating ranges are summarised in Table 2

Heating rate during pre-heat	Typical 1-2, max 4°C / sec
Final pre-heat temperature	Within 125°C of solder temp
Solder temperature	Max 235°C*
Cooling rate	Max 5°C / sec

*Maximum time of 10 seconds for double wave.

Table 2
Wave (Flow) Soldering Parameters.

Notes

ZETEX

Application Note 26
Issue 1 April 1996

Fast Charging Batteries with Zetex High Current PNP Transistors and Benchmarq Controller ICs

Neil Chadderton

Introduction

The advances of digital technology and a waiting market have created a huge demand for portable products including cellular telephones, PDAs, laptop computers, CD players, and gaming systems. By their nature, all of these products must derive their energy from an integral rechargeable cell/battery pack that must periodally be re-charged from a mains outlet. The type of battery employed, or its chemistry, depends on the overall pricing available to the product - this will depend on the perceived product lifetime, the required energy/weight ratio, and which marketable advantages there may be apparent from reducing the number of re-charge cycles.

For the latest generations of products, users are expecting higher performance, less re-charge cycles, longer battery pack life and faster charge times. All of these parameters are very dependent on the manner in which the cell or battery pack is treated during the charging process. This includes consideration of the preferred cell charging method, thermal concerns, and pre-charge conditions.

Fast Charge Controller ICs

There are a number of fast charge controller ICs available that perform all the necessary monitoring and regulation control functions on one monolithic IC. The Benchmarq Microelectronics series of fast charge controller ICs monitor the terminal voltage to determine state of charge, and also the temperature of the battery pack (via a thermistor) to prevent deterioration of the separator material. Fast charge termination is effected by any of the following:

- Delta temperature/delta time ($\Delta T/\Delta t$)
- Negative delta voltage ($-\Delta V$)
- Maximum temperature
- Maximum time
- Maximum voltage

To allow cost effective IC designs, typical fast charge controller ICs are manufactured on existing CMOS processes, that do not readily lend themselves to incorporating on-chip pass elements within system cost constraints. Therefore the IC controls an external discrete transistor that is used within a linear regulator, or a step-down frequency modulated DC-DC converter topology to provide a suitable current source for charging.

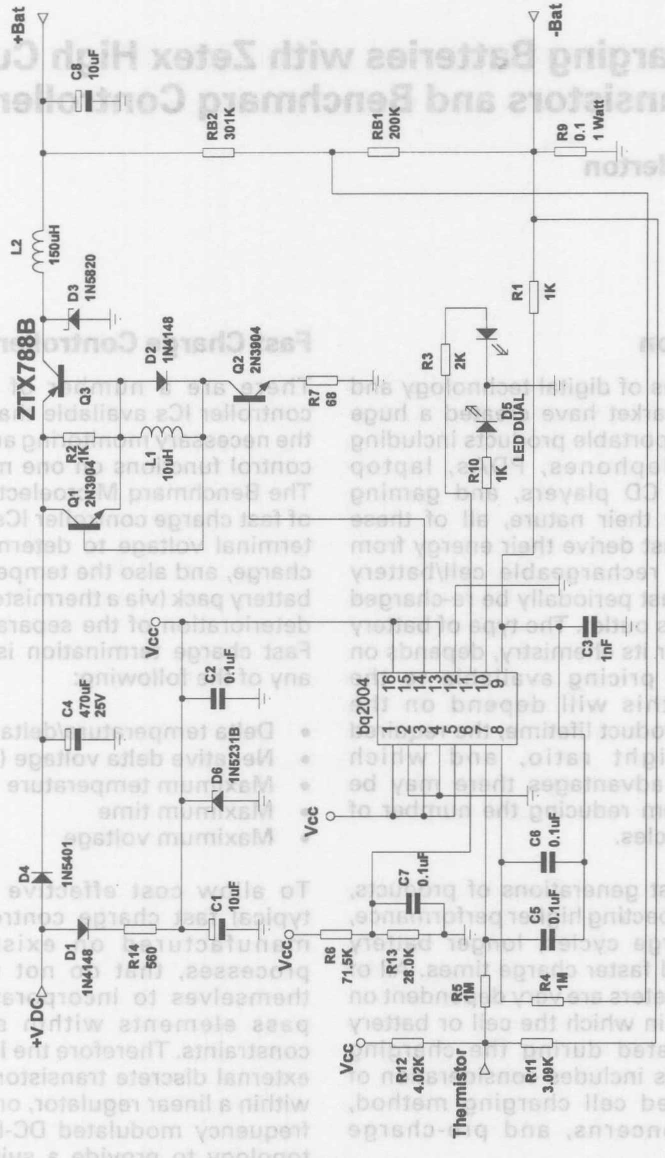


Figure 1
Fast Charge Circuit (Benchmarq Microelectronics) using the ZTX788B.

In general, switched mode circuits are preferred due to the higher efficiencies possible, and the smaller packaged pass devices required to perform the sourcing function. For example, a comparison of the power losses for a four 'C' cell battery pack charged from a 12V DC source at 2A, leads to values of greater than 12W for the linear control system, but less than 2W for the switched mode option.

The external pass transistor used within either circuit mode can be either a PNP Bipolar or a P-Channel MOSFET. For the linear regulator mode, the device must be capable of being heatsunk to remove the power dissipated, which will probably dictate that the product will be a large packaged, large silicon area device which will define a costly component. For the switched mode option, there is more choice available. Due to the higher silicon efficiency of the Bipolar technology compared with MOSFETs, high current fast switching PNP transistors are available that use a fraction of the silicon area required for comparably specified MOSFETs. This allows the bipolar chip option to be encapsulated within a smaller package, and therefore presents a lower cost to the designer.

The circuit shown in Figure 1 is one example of a circuit topology developed by Benchmarq using Zetex E-Line PNP transistors. The circuit allows the fast charging of Ni-Cd or Ni-MH battery packs for laptop computers and similar powered portable systems at a charge current of 2.3A, (defined by the current sense resistor R9). The potential divider formed by RB1 and RB2 is used to feedback a cell corrected level to the IC voltage sense input, that is then used to define the battery pack voltage. Other ICs/circuits are available from Benchmarq for higher current charging and for other cell chemistries.

The PNP transistor chosen for this particular application was the ZTX788B - one member of the ZTX788B to ZTX796A series. These devices are manufactured using a high gain emitter process to produce a Super-β h_{FE} range. The ZTX788B is a 15V, 3A continuous TO92 compatible part that exhibits very low $V_{CE(sat)}$, and can therefore replace much larger packaged components (such as TO220, TO126 and D-Pak) in switching applications. Figure 2 demonstrates the on-state voltage of the '788B as a function of load current with base drive current as parameter. Tables 1, 2 and 3 (Appendix A) summarise the pertinent parameters for the ZTX788B and other devices of interest for this application.

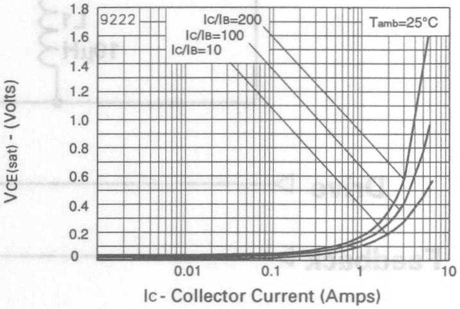


Figure 2
$V_{CE(sat)}$ v Collector Current Chart for the ZTX788B.

The circuit employs a circuit modification to provide an active turn-off for the transistor. [This doesn't however impact on the cost advantage over the MOSFET alternative]. Figure 3 shows the area around the pass transistor with the relevant level shift and turn-off components. The small signal switching transistor Q2, provides level translation from the "MOD" output of the controller IC, and its emitter resistor defines the base current for Q3, the ZTX788B. The inductor L2, Schottky diode (1N5820), and the 10µF capacitor are the Buck converter (step-down) components. The 1k base emitter resistor for Q3 would, in the absence of Q1, L1 and the signal diode (1N4148), provide passive turn-off for the pass transistor.

Figures 4 and 5 help to demonstrate the difference in turn-off performance for the two methods. These waveforms were recorded from a charger circuit using the bq2004 and ZTX789A for a 5 cell battery (output voltage to charge to 6V), and an input voltage of 12V. Both figures show the output from the IC controller (from the "MOD" pin), and the voltage from the collector of the ZTX789A with respect to ground.

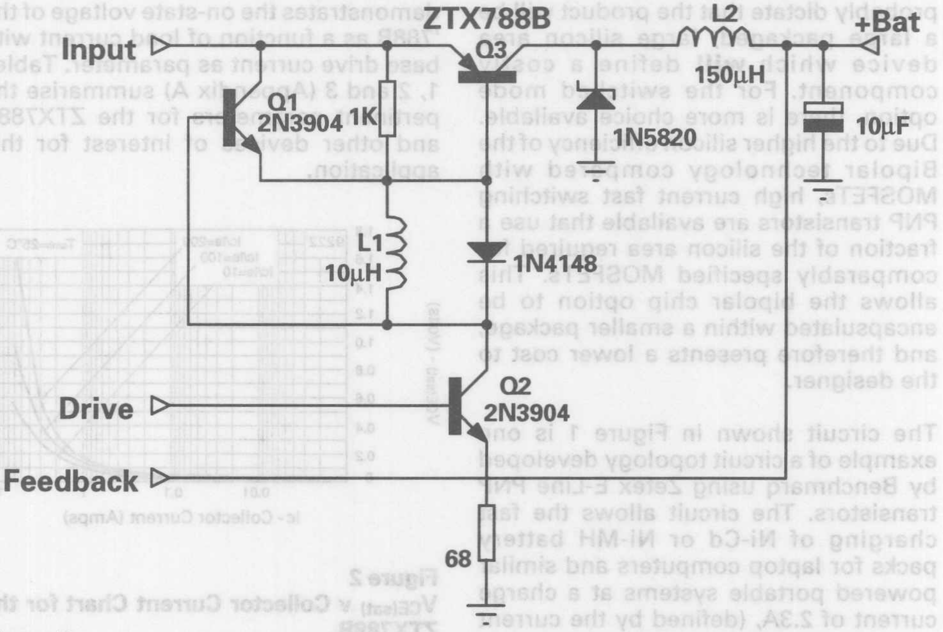


Figure 3
Active Turn-off circuit for Bipolar Transistors, allowing High Efficiency DC-DC Conversion at High Frequency.

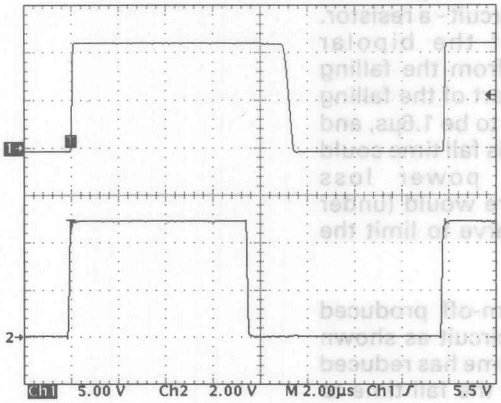


Figure 4
Oscillograph of Charger Circuit Employing Passive Turn-off for ZTX789A, showing: 1. Collector-to-0V waveform, and 2. IC "MOD" drive waveform. Note bipolar transistor storage and fall times. Channel 1: 5V/div, Channel 2, 2V/div, timebase at 2µs/div.

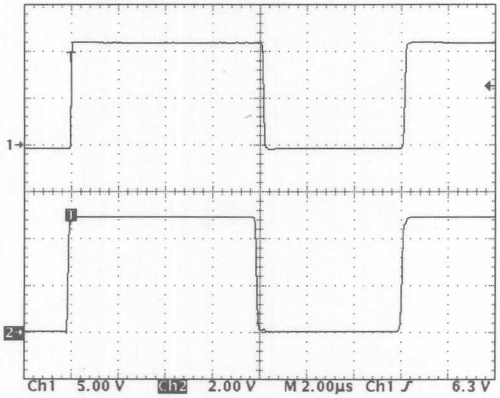


Figure 5
Oscillograph of Charger Circuit Employing Active Turn-off for ZTX789A, showing: 1. Collector-to-0v waveform, and 2. IC "MOD" drive waveform. Note bipolar transistor storage and fall times. Channel 1: 5V/div, Channel 2, 2V/div, timebase at 2µs/div.

Figure 4 shows the turn-off produced with a passive turn-off circuit - a resistor. The storage time of the bipolar transistor, or the time from the falling edge of trace 2 to the start of the falling edge of trace 1 is shown to be 1.6µs, and the fall time is 340ns. This fall time could represent a major power loss contributer and therefore would (under some circumstances) serve to limit the maximum load current.

Figure 5 shows the turn-off produced with an active turn-off circuit as shown in figure 3. The storage time has reduced to less than 200ns, and the fall time to 90ns. These switching values are comparable to, or better than the large P-Channel MOSFETs that would otherwise be specified, and allow the bipolar device to be operated to much higher currents than possible with passive turn-off circuitry.

Appendix A

Device	BV_{CEO}	$I_{C(DC)}$	I_{CM}	h_{FE}	$@I_C/V_{CE}$	$V_{CE(sat)}$ (max.)	$@I_C/I_B$	Package
FMMT718	20	1.5	6	230(typ)	2A/2V	145mV	1.5A/50mA	SOT23
FMMT720	40	1.5	4	290(typ)	1A/2V	150mV	1A/50mA	SOT23

Table 1. PNP Transistors for Fast Chargers <1A.
(Selected from the Zetex SuperSOT series. The FMMT718 chip is also available in the E-Line (enhanced TO92) package).

Device	BV_{CEO}	$I_{C(DC)}$	I_{CM}	h_{FE}	$@I_C/V_{CE}$	$V_{CE(sat)}$ (max.)	$@I_C/I_B$	Package
ZTX788B	15	3	8	300 min	2A/2V	450mV	2A/10mA	E-Line
ZTX789A	25	3	8	200 min	2A/2V	450mV	2A/20mA	E-Line
ZTX790A	40	2	6	200 min	1A/2V	450mV	1A/10mA	E-Line
ZTX949	30	4.5	20	140(typ)	5A/1V	320mV	5A/300mA	E-Line
ZTX951	60	4	15	120(typ)	4A/1V	300mV	4A/400mA	E-Line

Table 2. PNP Transistors for Fast Chargers 1 to 4A.
(Selected from the ZTX788B - ZTX976A, and ZTX950 series. These devices are available in the SOT223 surface mount package - prefixed FZT eg FZT788B).

Device	BV_{CEO}	$I_{C(DC)}$	I_{CM}	h_{FE}	$@I_C/V_{CE}$	$V_{CE(sat)}$ (max.)	$@I_C/I_B$	Package
ZBD949	30	5	20	140(typ)	5A/1V	450mV	5A/300mA	TO126
ZBD953	100	4	10	90(typ)	3A/1V	470mV	4A/400mA	TO126

Table 3. PNP Transistors for Linear and Switched Mode High Current (5A) Fast Chargers.

Notes

ZETEX

Application Note 27
Issue 1 June 1996

ZR431 Application Note

David Bradbury

The ZR431 is an enhanced version of the industry standard 431. It is a three terminal shunt regulator giving excellent temperature stability and the capability of operating at currents from 50µA up to100mA. Its output voltage can be set to any voltage in the range of Vref (2.5V) to 20V by the addition of two external divider resistors. The reference input current is typically only 100nA, so high value resistors can be used without error. Its minimum operating current and reference input current are ten times lower than industry standard 431 parts. (Please refer to Appendix A)

Following are a few example applications of the ZR431 which show how the parts can be used.

Shunt Regulator

The ZR431 is ideal for providing low voltage stabilised supplies or references. Simple low output voltage supplies can be difficult to build as the components normally used, such as zener diodes or band gap references, are either too poor in performance or not available at the desired voltage (band-gap devices are usually fixed at 2.5V or 5V with no intermediate values).

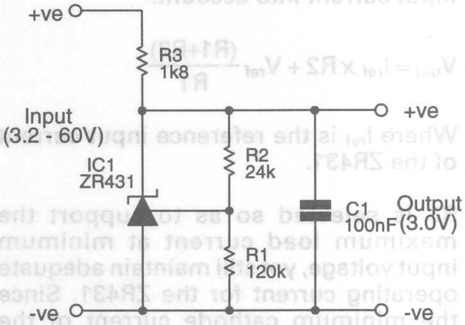


Figure 1
ZR431 used as a 3V Shunt Regulator.

The circuit in Figure 1 shows a 3V shunt regulator utilising the ZR431. The output of this supply can be set to any voltage in the range 2.5V to 20V by adjusting the ratio of R1 and R2, following the relationship:-

$$V_{out} = V_{ref}\,\frac{(R1+R2)}{R1}$$

Where V_{ref} is the reference voltage of the ZR431.

Note that a small correction to this formula is required if R1 and R2 are given very high values, since the reference input current of the ZR431 flows through R2 but not R1 thereby causing a small error. For the Zetex ZR431, this reference current is 200nA maximum so as long as a current of at least 20µA is passed through R1, setting its maximum value at 60k ohms, the reference current can be ignored. For very low current applications where it is advantageous to set R1 and R2 at a level where the reference current must be considered, the following formula takes input current into account:-

$$V_{out} = I_{ref} \times R2 + V_{ref} \frac{(R1+R2)}{R1}$$

Where I_{ref} is the reference input current of the ZR431.

R3 is selected so as to support the maximum load current at minimum input voltage, yet still maintain adequate operating current for the ZR431. Since the minimum cathode current of the ZR431 is only 50µA, it rarely has an effect on the choice of R3.

$$R3 = \frac{V_{in_{(min)}} - V_{out}}{I_{out_{(max)}} + I_{z(min)}}$$

Where $I_{z(min)}$ is the minimum cathode current of the ZR431.

The low series resistance of the ZR431 not only gives the shunt regulator good load regulation but also a line rejection of over 60dB. In this circuit, capacitor C1 both maintains this rejection figure at high frequencies and ensures stability should the power supply's load be capacitive. For the Zetex ZR431, a capacitor greater than 1.5nF will ensure stability for any load. (Competitors' versions of the '431 generally require values of 3.3µF or greater).

Switch-Mode Power Supply Controller

The circuit shown in Figure 2 is commonly used in the control loop of switch-mode power supplies. The output voltage is sensed via R3/R4 by the ZR431, and the IC controls the current passed through an opto-coupler and hence feeds back output voltage status to the switching regulator. The minimum supply voltage on which this popular circuit can be used is set by the minimum cathode voltage of the regulator IC, plus the forward voltage drop of the opto-coupler LED. This limits the minimum operating voltage of the circuit to around 4V.

The components C1 and R2 are not required for correct operation of the ZR431. They have been included as they are frequently needed to stabilise the overall control loop of the switch-mode power supply of which Figure 2 is only a part. **The ZR431 is unconditionally stable without additional components.**

Note 1:
Resistor R5 is only necessary if the minimum operating current of the ZR431 causes excess opto-coupler current before the reference functions correctly. The minimum operating current of the ZR431 is only 50µA so R5 is rarely required when using the Zetex part. (Competitors versions can pass in excess of 1mA before operating correctly, hence the occasional need for R5).

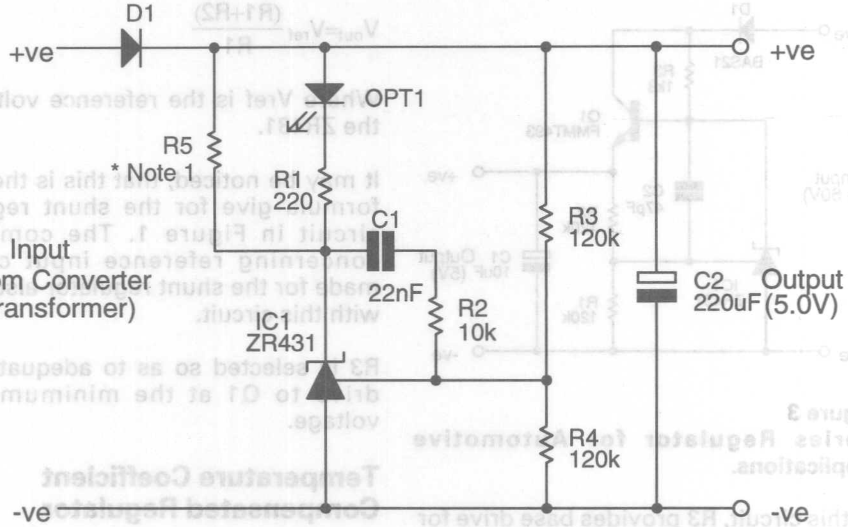


Figure 2
Control Loop of a Switched Mode Power Supply : C1, R2 and R5 are Optional Components.

Series Regulator Power Supply

Although shunt regulators such as the circuit given in Figure 1 provide very simple and low cost power supplies, they are often inefficient because the pass resistor R3 must be selected to supply the maximum load expected at the minimum input voltage. When the actual load is lower or the input voltage is higher than for these worst-case conditions, excess current must be shunted away by the ZR431. This is not usually a problem for low output current supplies or when efficiency is not a major concern, but for loads greater than a few milliamps losses can be significant.

However, a series regulator can significantly reduce power losses and hence improve efficiencies. Series regulators are usually much more complex than shunt circuits, requiring a reference, error amplifier, driver and series pass elements. However, it is possible to construct a very simple series regulator using a ZR431 to perform all but the pass transistor function. Figure 3 shows a series regulator for an automotive application. The circuit has been designed to provide 30mA at 5V to a microcontroller, to operate with a normal input supply range of 7V-15V, to withstand load-dump supply transients of 60V and a sustained overvoltage input of 24V, and tolerate reverse battery connection. All parts used are surface-mount so the supply can be constructed using little PCB area.

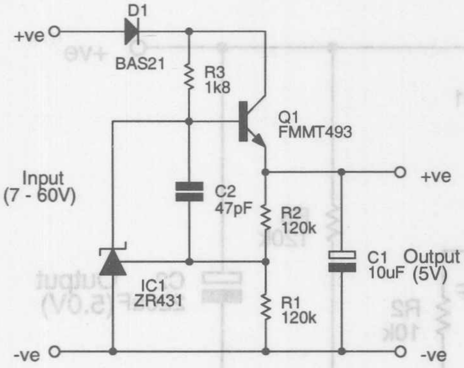


Figure 3
Series Regulator for Automotive Applications.

In this circuit, R3 provides base drive for the series pass transistor Q1. The ZR431 senses the output voltage of the supply via R1/R2, compares this with its internal reference and shunts excess base drive from Q1 so as to maintain the required supply output. Reverse polarity protection is provided by D1. The output resistance of the supply is around 20mΩ. Capacitor C1 helps maintain this very low output impedance at high frequencies. The stabilising capacitor C2 has been kept small so as not to degrade the excellent high frequency performance of the ZR431. The circuit will supply a 30mA load at minimum input and for higher inputs it can source much more. Note that the circuit does not include a current limit, and so care must be taken not to short the output for sustained periods.

The output of this type of supply can be set to any voltage in the range 2.5V to 20V by adjusting the ratio of R1 and R2, following the relationship:-

$$V_{out}=V_{ref}\frac{(R1+R2)}{R1}$$

Where Vref is the reference voltage of the ZR431.

It may be noticed, that this is the same formula give for the shunt regulator circuit in Figure 1. The comments concerning reference input current made for the shunt regulator also apply with this circuit.

R3 is selected so as to adequate base drive to Q1 at the minimum input voltage.

Temperature Coefficient Compensated Regulator

There are often occasions when a reference with a large and closely defined temperature coefficient (T.C.) are required. A common example of this is in the management of rechargeable batteries. Certain cell technologies such as Lithium and Lead-acid require charging to a set voltage to ensure they are fully charged. Failure to do this adequately can result in a reduction of battery life and in extreme cases, cell rupture. Unfortunately, this end of charge voltage varies with temperature and so the voltage reference controlling charge termination must be temperature compensated. Following is an example of how an accurate, low temperature coefficient regulator can be modified to produce a temperature compensated reference for lead-acid battery charging applications.

By adding a transistor with a known temperature coefficient to the reference input circuit of the ZR431, a reference

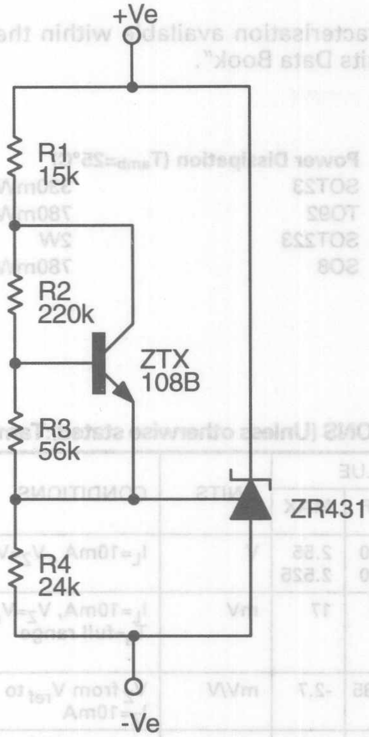


Figure 4
Temperature Coefficient Compensated Regulator.

with the same T.C. as the lead-acid batteries can be produced. Figure 4 shows a 6.9V shunt reference which has a T.C. of -11.7mV/°C, matching a three cell lead-acid battery pack terminal voltage and temperature coefficient.

In this T.C. corrected reference circuit, the ratio of R2/R3 sets the overall temperature coefficient by amplifying the well characterised negative temperature coefficient of the ZTX108B Vbe. (The ratio will require modification should another transistor type be substituted). The ratio of R1/R4 has been adjusted to set the final output voltage to 6.9V at 25°C. The exceptionally low bias current of the ZR431 means that no allowance need be taken of the regulators reference input current in this circuit. Thanks to the low quiescent current of the ZR431, the shunt reference circuit will operate at currents down to less than 200µA. Note, for best performance, the reference circuit should be situated close to the battery pack under charge to ensure temperature tracking.

Appendix A

Partial Characterisation of ZR431. Full Characterisation available within the "High Performance Linear Bipolar Integrated Circuits Data Book".

ABSOLUTE MAXIMUM RATING

Cathode Voltage (V_Z)	20V
Cathode Current	150mA
Operating Temperature	-40 to 85°C
Storage Temperature	-55 to 125°C

Power Dissipation (T_{amb}=25°C)

SOT23	330mW
TO92	780mW
SOT223	2W
SO8	780mW

Recommended Operating Conditions

	Min	Max
Cathode Voltage	Vref	20V
Cathode Current	50µA	100mA

ELECTRICAL CHARACTERISTICS TEST CONDITIONS (Unless otherwise stated):Tamb=25°C

PARAMETER	SYMBOL	VALUE			UNITS	CONDITIONS
		MIN	TYP	MAX		
Reference Voltage 2% 1%	V_{ref}	2.45 2.475	2.50 2.50	2.55 2.525	V	I_L=10mA , V_Z=V_{ref}
Deviation of Reference Input Voltage over Temperature	V_{dev}		8.0	17	mV	I_L=10mA, V_Z=V_{ref} T_a=full range
Ratio of the change in Reference Voltage to the Change in Cathode Voltage	$\frac{\Delta V_{ref}}{\Delta V_Z}$		-1.85	-2.7	mV/V	V_Z from V_{ref} to 10V I_Z=10mA
			-1.0	-2.0	mV/V	V_Z from 10V to 20V I_Z=10mA
Reference Input Current	I_{ref}		0.12	1.0	µA	R1=10K, R2=O/C, I_L=10mA
Deviation of Reference Input Current over Temperature	ΔI_{ref}		0.04	0.2	µA	R1=10K, R2=O/C, I_L=10mA T_a=full range
Minimum Cathode Current for Regulation	I_{Zmin}		35	50	µA	V_Z=V_{ref}
Off-state Current	I_{Zoff}			0.1	µA	V_Z=20V, V_{ref}=0V
Dynamic Output Impedance	R_Z			0.75	Ω	V_Z=V_{ref} f=0Hz

ZETEX

Application Note 28
Issue 1 June 1996

The ZLDO series of Low Dropout Voltage Regulators

David Bradbury

Introduction

The ZLDO series of low dropout linear regulators operate with an exceptionally low dropout voltage, typically only 30mV with a load current of 100mA. The series feature output voltages in the range 3 to 18V, supply loads up to 300mA, and yet consumes typical quiescent currents of only 630µA. The parts include a logic level shutdown control and provide a low battery warning flag. Designed for where power losses must be minimised for efficiency or thermal reasons, the ZLDO Series regulators have many potential uses.

Operation From A Low Voltage Battery Pack

Figure 1 shows the ZLDO500 regulator being used to stabilise the output of a 6V battery pack. The ultra low dropout voltage of only 100mV at full load (300mA) given by the regulator, allows the minimum number of cells to be used in the pack, and also maximises the energy that can be removed from the battery before the output of the regulator starts to fail. At a load current of 100mA the dropout voltage falls to around 30mV. The endurance of the battery pack is not only dependent on dropout voltage. When operating, some low dropout regulators can consume high quiescent currents, sometimes approaching as much as a tenth of their maximum load current specification when approaching dropout conditions. Despite its 300mA output rating, when enabled the ZLDO series of devices consume typically only 600µA when regulating normally, and 3mA when the input falls too low for regulation.

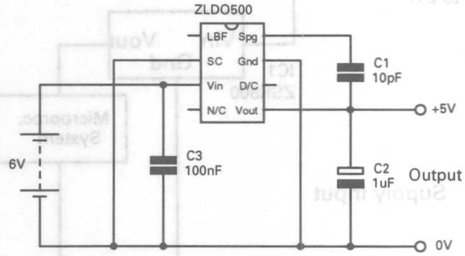


Figure 1
Basic ZLDO500 Based Battery Powered Supply.

Distributed Power Supplies

A common problem with large multiple board logic systems is that the total supply current taken at 5V can become excessive, causing voltage drops and noise in the power supply wiring unless heavy cables and large decoupling capacitors are used. A convenient solution to this problem is to provide power using a higher voltage supply locally regulated to 5V. Voltage drops and noise are now eliminated by the regulators but they introduce a new problem of significantly increased power losses if standard regulators are used. By employing the ZLDO500 for the local regulators in a circuit similar to Figure 1 but repeated on each logic board, the power supply to the logic boards can be distributed at a voltage close to 5V. This will largely eliminate the added losses of a distributed power supply system, whilst minimising the supply voltage errors and noise.

Logic Controlled Power Supply

Figure 2 shows all that is necessary to allow a microprocessor to control a power supply based on the ZLDO500. The Shutdown Control pin (pin 2), is a logic compatible input that disables the regulator when a voltage in excess of 1.5V is applied. The current required to drive this input is less than 10µA. When the regulator is shutdown in this way, the quiescent current of the ZLDO500 falls to around 10µA. This makes the regulator suitable for a wide range of battery powered applications where intermittent operation occurs. The shutdown control pin should not be taken to a voltage higher than Vin if low quiescent supply current is important. The shutdown control is a high impedance input and so if not required, should be wired to the ground pin (pin 7).

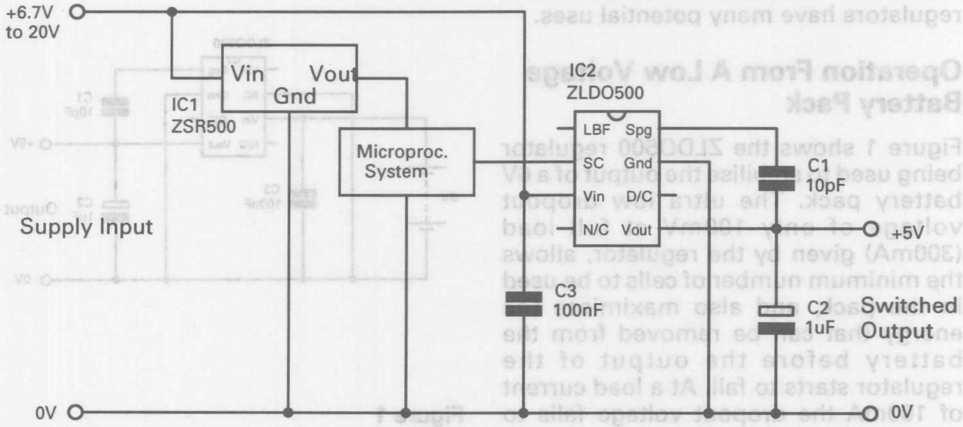


Figure 2
Employing the Shutdown Pin to Conserve Battery Capacity.

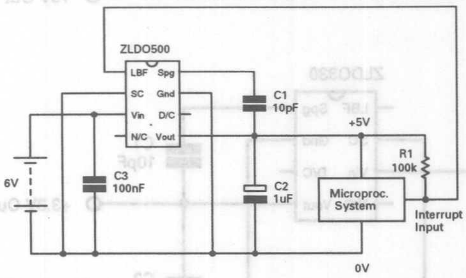


Figure 3
Use of the Low Battery Flag (LBF) to

Low Battery Flag

The ZLDO500 series provides an output called Low Battery Flag (LBF). Unlike many regulators that only signal that they are falling out of regulation, the LBF output of the ZLDO series indicates that the voltage drop across the regulator has fallen to less than 400mV and so supply failure is imminent. This improved warning gives both more time for the system to shutdown gracefully, and maintains regulation while this happens. This could be a vital point for instance if measurements are under way and must be completed accurately. The LBF output is driven by an open collector NPN transistor which pulls low when the supply to the regulator is failing. Figure 3 shows this output being used. Note that resistor R1 is necessary only if the interrupt logic does not include a pull-up resistor.

Simple 3.3V Supply

Using a circuit such as Figure 4, the ZLDO330 can easily provide a 3.3V logic supply from an available 5V rail where most standard regulators could not guarantee correct operation. Although this approach is not particularly energy efficient, if the load taken at 3.3V is not too large, then the added complexity and cost of a 3.3V switching converter may not be justifiable and so this linear solution can be preferable. This circuit will also give far less noise than a switching regulator which can be important when handling low level analogue signals or low voltage measurements.

Post Converter Regulation

A common problem with multiple output switch mode converters is that only one output can be used in the feedback control loop of the switching regulator. Thus only one output is fully regulated. All other outputs are prone to tracking errors that occur if the load on any output change significantly. By ensuring close coupling of all transformer windings and minimising the impedance of all outputs, these errors can be reduced but never eliminated. A simple solution to this problem is to wind the switching

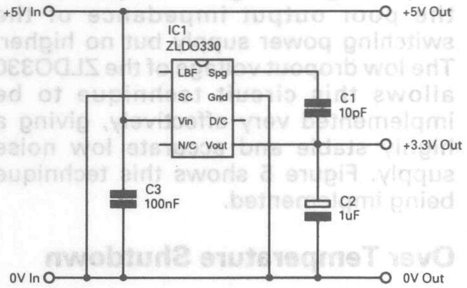


Figure 4
ZLDO330 Used to Regulate 5V to 3.3V.

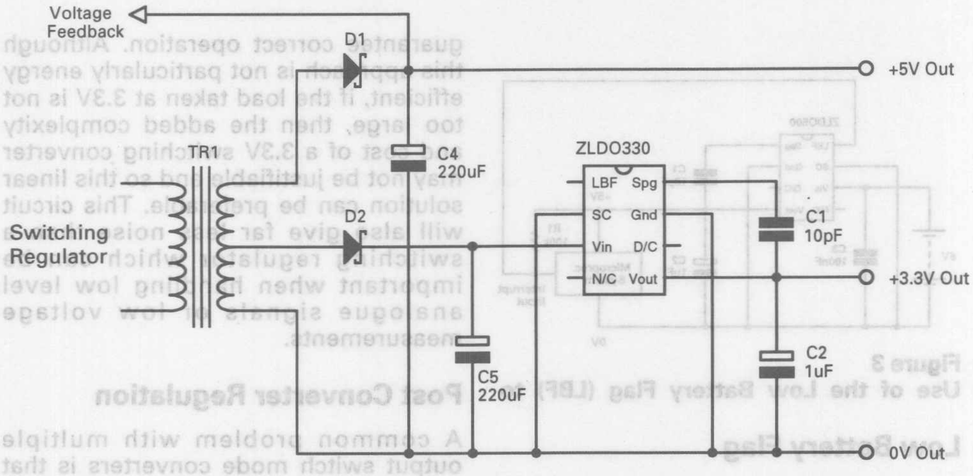


Figure 5
ZLDO Switched Mode Supply Post-Regulator to Improve Output Impedance and Noise Performance.

regulator transformer to give a slightly higher voltage than required and regulate down from this to the desired voltage with a linear regulator. To keep losses low and so maintain the advantages of a switch mode supply, it is important that the voltage drop across this regulator is kept as low as possible, i.e. just high enough to compensate for the poor output impedance of the switching power supply but no higher. The low dropout voltage of the ZLDO330 allows this circuit technique to be implemented very effectively, giving a highly stable and accurate low noise supply. Figure 5 shows this technique being implemented.

Over Temperature Shutdown

The ZLDO regulator series include an over temperature shutdown circuit that disables the regulator if its chip temperature should exceed 125°C for any reason. Although intended to provide a limited guard against excessive internal power dissipation, this circuit will shut down the regulator if the ambient rises above 125°C. Thus, the regulator could be used to disable a circuit in the event of the ambient temperature within which the circuit is mounted becoming too high. Any internal power dissipation due to load current will reduce the ambient temperature at which shutdown occurs to some extent. A consequence of achieving the extremely low dropout voltage and high current performance provided by the ZLDO series, is that the parts can be damaged by sustained output shorts or excessive loads when combined with high input supply voltages. To ensure reliable operation, keep loads within the SOA graph boundaries provided on the data sheet for the respective part.

Section 2

Design Notes

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ZETEX

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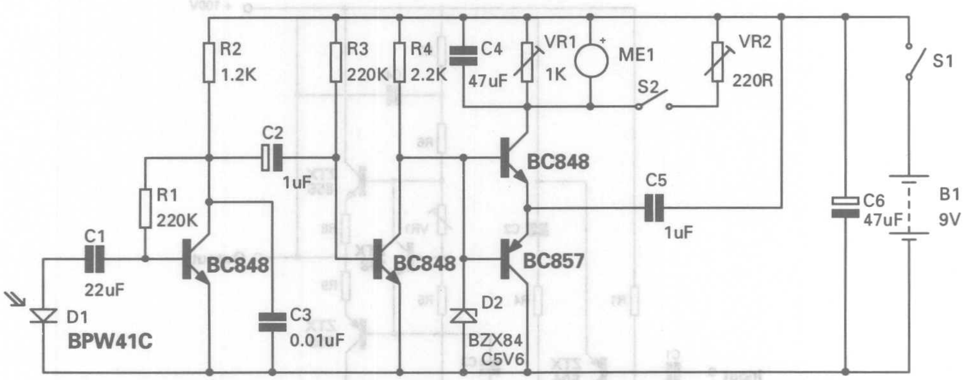
Design Note 1
Issue 2 June 1995

Light Coupled Speed Measurement Using the BPW41 Photodiode



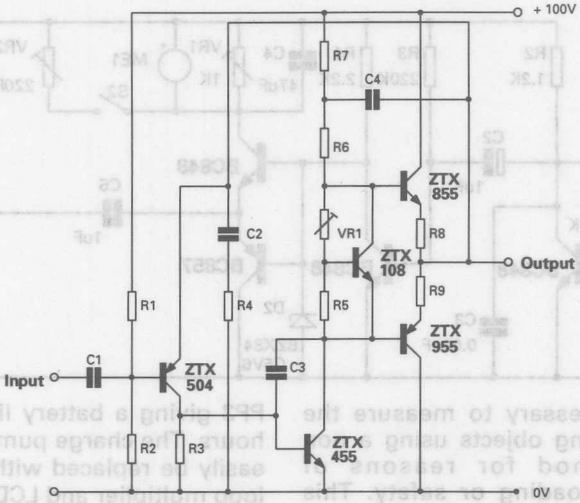
It is often necessary to measure the speed of rotating objects using a non contact method for reasons of convenience, loading or safety. This simple circuit provides flicker frequency measurement to enable the speed of a rotating object to be obtained.

The circuit uses the inexpensive BPW41C photodiode in photovoltaic mode enabling operation in light levels from 200 lux up to direct sunlight. Employing a signal and limiting amplifier, a low temperature coefficient zener controlled charge pump circuit and an analog meter for speed display, the unit draws around 10mA from a 9V PP3 giving a battery life of at least 50 hours. The charge pump and meter can easily be replaced with a phase locked loop multiplier and LCD pulse counter if a more accurate digital display is required. Low cost small signal surface mount transistors were employed in the original design, but could be substituted with ZTX300/ZTX500 parts for the NPN/PNP positions respectively for a through-hole design.

Originally intended for measuring the RPM of small engines for tuning purposes, the circuit could easily be adapted for a wide range of industrial speed measurement tasks too.

Design Note 2
Issue 2 June 1995

Sonar Transducer Driver using TO92 Style Bipolar Transistors



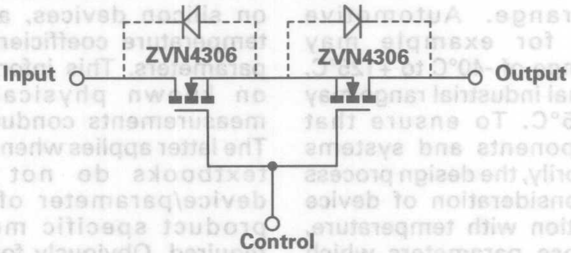
Depth/Echo sounders require high power linear amplifiers for driving the transducer. Since these amplifiers operate in pulsed mode, the average power dissipation experienced by the output transistors is low, but the value of the pulsed current can be very high. This would normally dictate that high power, large package power transistors would be required, which tends to increase the overall circuit size.

With a peak current capability of 10A, the ZTX855/ZTX955 transistors are packaged in the TO92 style E-Line package, and permit an output power of 100W to be obtained from this design. Larger and more costly TO220 transistors have been eliminated. The devices possess an F_T figure of over 90MHz, enabling a high bandwidth and slew rate specification.

	BV_{CBO}	BV_{CEO}	IC(M)	I_C (DC)	Package
ZTX855	250V	150V	10A	4A	TO92
ZTX955	180V	140V	10A	3A	TO92

Design Note 3
Issue 2 June 1995

High Current High Voltage Analog Switches



Commercial analog switch ICs are in the main designed for low power signal applications, and often cannot be used due to:

a) not possessing a wide enough swing in signal voltage;

b) size problems;

c) current limitations due to their on-state resistance.

Single MOSFET devices cannot be used as alternatives because their body diode (shown dotted in the diagram) will conduct part of any AC signal even when the device is turned off. However two low resistance ZVN4306A TO92 style MOSFETs in the simple arrangement shown here, will give a compact high performance 0.45Ω, 1A analog switch without this restriction. With the control input driven 2V more positive than the peak of the analog signal to turn the switch on, and as negative to turn the switch off, the circuit above will accommodate signals up to 18V peak to peak. Substituting Zetex 2N7000 MOSFETs in the circuit allows signals up to 38V ptp and still handles currents up to 200mA. Higher voltage signals may be switched using isolated drive circuits.

For applications where size is critical, high performance Zetex SOTFETs can be used. The SOT23 ZVN4106F will produce an analog switch possessing an on-resistance of only 4Ω.

Temperature Effects On Silicon Semiconductor Devices

Many applications require reliable circuit operation over a wide temperature range. Automotive environments for example may experience a range of -40°C to +125°C, while a more usual industrial range may be -10°C to 85°C. To ensure that electronic components and systems behave satisfactorily, the design process must include consideration of device parameter variation with temperature, in particular those parameters which affect leakage currents, drive conditions, on-state losses, and switching performance.

This design note provides a general indication of the effects of temperature on silicon devices, and lists typical temperature coefficients for the main parameters. This information is based on known physical effects, and measurements conducted previously. The latter applies when either the device textbooks do not consider the device/parameter of interest, or a product specific measurement is required. Obviously for such a general guide, these values must be considered as approximate, but will probably be adequate in many cases.

BIPOLAR (except Darlingtons)

Parameter	Temp. Coefficient	Comment
h_{FE}	+0.3 to 0.6%/°C	In active region at low to medium currents
	Negative	At high currents. Cross over point falls as I_C and Breakdown Voltage Increases
	-0.25%/°C	In saturation region
$V_{BE(sat)}$	-0.15%/°C	
$V_{BE(on)}$	-1.9 to 2.2 mV/°C	PNP tend to be higher than NPN
$V_{CE(sat)}$	+0.23 to 0.4%/°C	Small signal devices (eg ZTX300)
	0.12 to 0.15%/°C	Medium power devices (eg ZTX650)
$V_{CBO}(V_{CES})$	+0.07 to 0.09%/°C	
V_{CEO}	+ or - about zero	(Eg. can be small, but very device dependent)
V_{EBO}	+0.04%/°C	
I_{CBO}.	x2 per 11°C	
I_{CEO}	Approx x4 per11°C rise	Very dependent on h_{FE}
$I_{CER(10k)}$	Approx x3 per 11°C rise	
t_d and t_r	Independent	
t_s	+1.4 to 1.6%/°C	
t_f	+1.0 to 2.5%/°C	

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DARLINGTONS

Parameter	Temperature Coefficient	Comment
h_{FE}	+0.5 to 1.6%/°C	In active region at low to medium currents
	+ or - about zero (small) (due to inability to saturate fully)	In saturation region
$V_{CE(sat)}$	+ or - about zero (small)	
$V_{BE(on)}$	-0.2%/°C	
V_{CBO}	+0.08%/°C	
V_{EBO}	+0.04%/°C	
I_{CEO}	Approx x2 per 8-10 °C rise.	Dependent on voltage as well as temperature

MOSFETS

Parameter	Temperature Coefficient	Comment
V_{DSS}	+0.04 to 0.1%/°C	Increases with resistivity (BV_{DSS})
$R_{DS(ON)}$	+0.7 to 1.0%/°C	
$V_{GS(TH)}$	-0.1 to 0.2%/°C	
I_{DSS}	Approx x2 per 11°C rise	
g_{fs}	Small (say -0.2% / °C)	
V_f (of body diode)	-2 to 2.5mV/°C	
dV/dT capability falls as temperature increases. Switching speed does not alter significantly.		

DIODES

Parameter	Temperature Coefficient	Comment
V_{SS}	-1.9 to 2.2mV/°C	
B_{SS}	+0.7 to 0.09%/°C	

VARICAP/VARACTOR DIODES

Parameter	Temperature Coefficient	Comment
Capacitance	Abrupt:280 ppm @ 4V; 100 ppm @ 20V	Very dependent on bias voltage
	Hyperabrupt:350 ppm @ 3V; 80 ppm @ 20V	

VOLTAGE REFERENCE DIODES

Parameter	Temperature Coefficient	Comment
V_Z	-0.06%/°C	3.3V(Field effect)
	Nominally zero	5.0V
	+0.075%/°C	12.0V (Avalanche effect)
	0.2%/°C	24.0V (Avalanche effect)

SCHOTTKY DIODES

Parameter	Temperature Coefficient	Comment
V_F	-1.2mV/°C	

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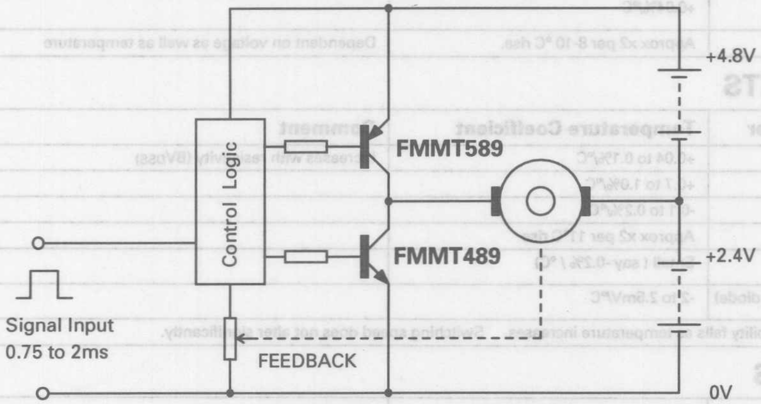
Remote Control Servo Amplifier using Protection Free 1A DC Rated SOT23 Devices



The Zetex high performance SOT23 package, permitting power dissipation to 500mW (now up to 625mW with the FMMT618 and FMMT718 "**SuperSOT**" series), and the high efficiency of the 'MATRIX' geometry design, produces a performance previously only available in TO92, SOT89 or larger package styles. This advantage can be demonstrated by the servo amplifier driver shown above.

This circuit uses the high gain 30V BV_{CEO} rated FMMT489 and FMMT589, that possess typical gains of 200 at 1A and 0.5A respectively, low $V_{CE(sat)}$ (0.3V and 0.35V at 1A), and capable of handling peak currents of up to 2A, and 1A continuous. These features allow surface mount designs to be realised at low cost, to exhibit better switching performance, and using a minimum of board area. Higher voltage variants are available as the FMMT491 and FMMT591, which are rated at 60V.

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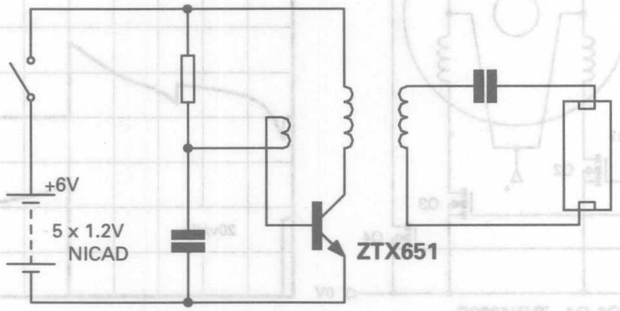
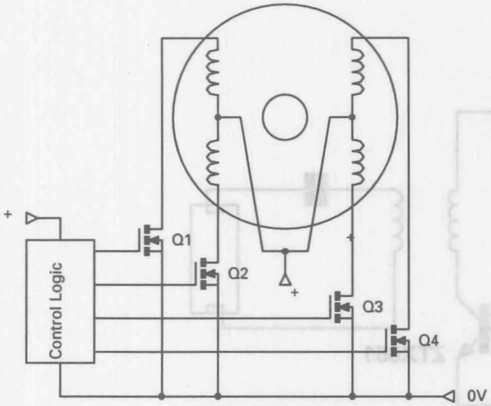
6V Battery Operated Fluorescent Lamp



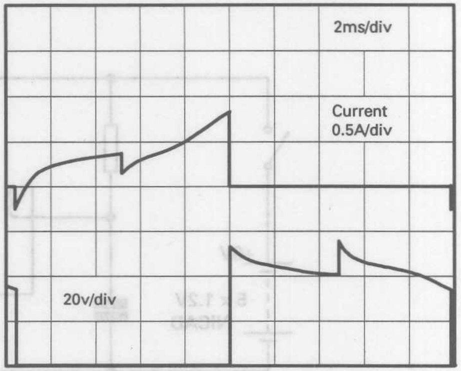
To handle tube striking and normal operating characteristics, the converter transistor used in this simple fluorescent lamp ballast requires a high current capability and a high breakdown voltage. The ZTX651, possessing a B_{VCBO} rating of 80V, and a current rating of 2A continuous and 6A peak, performs more efficiently than the expensive TO220 transistor normally used. Due to the lower on-state losses of the Zetex transistor, the circuit is capable of 75% efficiency, with a resulting 12% increase in battery life. With a saturation voltage of only 0.2V at 2A, power dissipation in the E-Line (TO92 style) transistor is kept well within its 1W rating. The small size of the ZTX651 and the converter's 25kHz oscillation frequency allows the circuit to be constructed in a compact form, small enough to fit in a torch handle, etc.

Higher gain variants (Super-β devices such as the ZTX688B and ZTX788B series) are now available to help reduce base drive requirements and thereby improve efficiency still further.

Printer Stepper Motor Drivers



Q1-Q4 ZVN4206C



Many modern printers employ a number of stepper motors to drive the various print and transport mechanisms.

Direct logic drive and high load currents prompt manufacturers to use expensive and bulky TO220 or TO126 Darlington transistors in these circuits. In many cases devices that are needlessly overrated for this application.

The ZVN4206C MOSFET is the perfect drop-in replacement for these parts. This centre drain TO92 style MOSFET will handle peak currents of up to 8A and being characterised for 5V gate drive, can be directly driven from most logic families.

The inherent body diode and BV_{DSS} of 60V permit operation without additional transient protection or dV/dt shaping.

The ZVN4206 is also available in SOT223 as the ZVN4206G.

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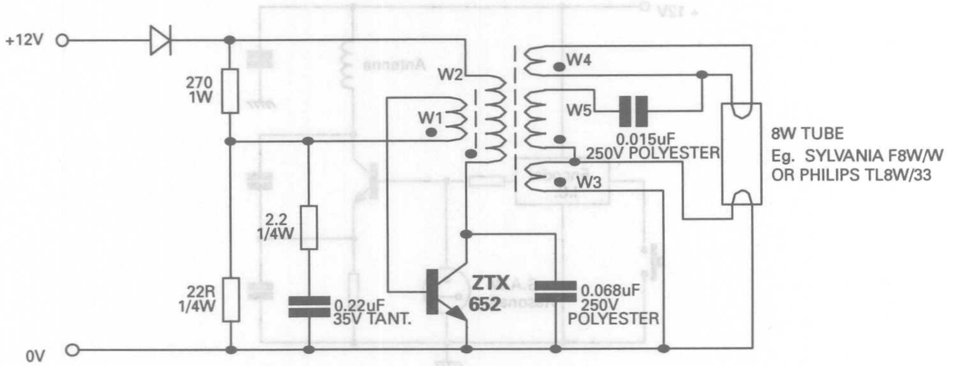
8W, 12V Fluorescent Lamp



Fluorescent tubes are notoriously difficult loads to drive. Before striking, the tube effectively appears open circuit, needing 300V or greater to provoke conduction. After striking they adopt a low voltage characteristic where the required current can be sustained at low voltage, typically 40V. These changes in characteristics place severe demands on any DC-AC converter and the associated transistors used to drive the tubes.

The combined characteristics of high current capability, low switching losses and high breakdown voltage of the TO92 style ZTX652 allows this simple automotive/ caravan/RV fluorescent lamp to be constructed without the size and cost penalties of the TO220 transistor normally needed for this application. Handling 90V pulses during striking, and 3A peak current during normal operation, the ZTX652 enables this circuit to run at 20kHz with an efficiency of over 80%.

The 1.5W PCB mounted power rating of the ZTX652 gives this converter the capability of operating safely from supplies in the range of 10-16.5V, even under fault conditions such as failed or disconnected fluorescent tubes.

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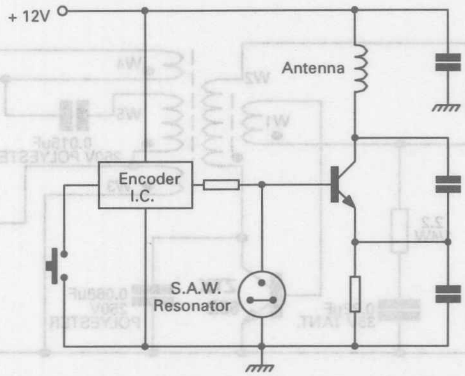
Automotive Security Systems and RF Transistor Products



The majority of modem automotive alarms employ coded RF systems that key a VHF carrier wave with a unique (or at least, one of many) code. The encoding ICs available are able to keep component count to a minimum, and for efficient, cost-effective and non-critical operation, a high performance RF driver is therefore essential.

The transmitter circuit shown, is typical of such systems. It requires a device with tightly controlled DC and RF parameters, as there is no provision for component variability.

World-wide, different frequencies have been allocated for this application and this defines device selection. For the UK where a band at 418MHz has been set, the ZTX325 with a peak F_T of 3.2GHz is ideal.The US band of 314MHz is well served by the Zetex MPSH10 with an F_T of 1.6GHz peak and various European frequencies in the range of 200-300MHz are covered by the ZTX321 with an F_T peaking at 880MHz.

All these devices are available in the compact E-Line (TO92 style -though much slimmer) package or for ultra miniature designs, in SOT23.

RF output power is normally regulation limited to 0.5 -1W. Supply currents of 10 - 20mA are necessary to provide these levels due to antenna shrouding by the operators hand. At these current levels careful attention must be paid to available RF gain, F_T profiles, and biasing. The aforementioned transistors have all been characterised in this respect, and 's' parameters are available for the BFS17 and MPSH10.

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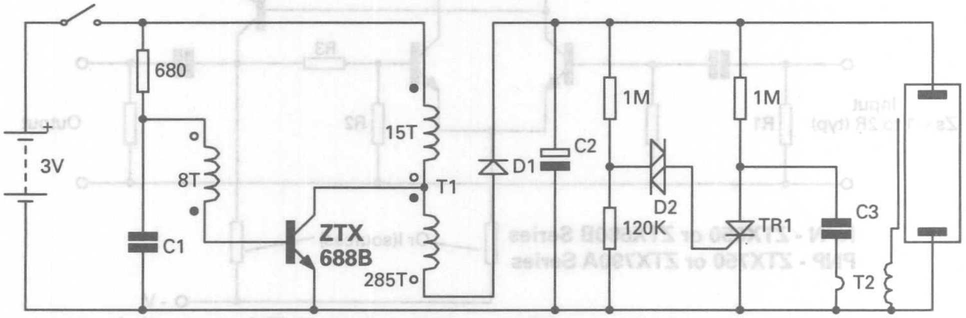
3V Distress Beacon

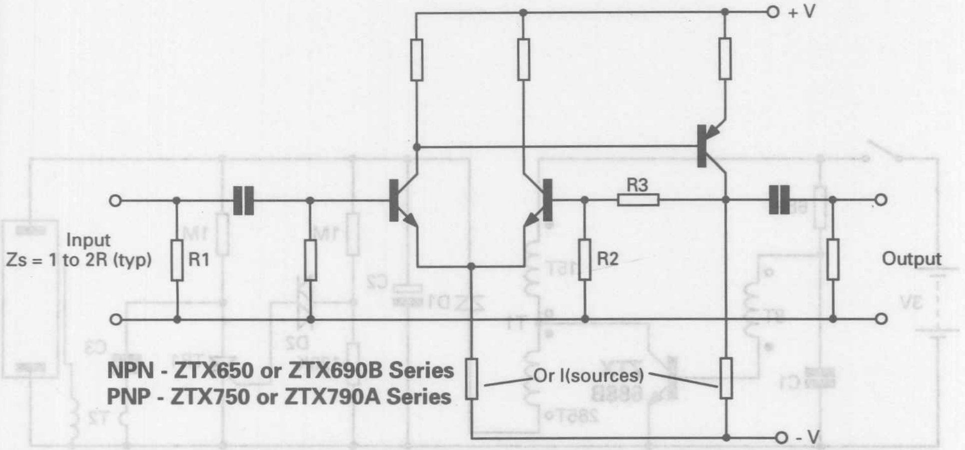


Essential characteristics of personal distress beacons (as used in hazardous areas by fire-fighters, mountain and air/sea rescue, etc.) are that they are compact, reliable and have good endurance capabilities. This leads to two cell supply designs whose efficiency is often degraded by the switching performance of the converter transistor.

A high gain low saturation voltage ZTX688B allows this 3V powered personal alarm beacon to operate with an efficiency of more than 70%, maximising flash brightness and battery life. This efficiency is obtained by combining an energy saving flyback design with the low on-state losses and low base drive requirements of the Zetex matrix based ZTX688B (which features a saturation voltage of 0.1V at 1A with only 5mA base drive). The small size of the E-Line (TO92 style) package helps to keep unit size to a minimum. Battery life is further extended by the design's capability of operating with supplies down to 1V.

Circuit action is as follows: The converter circuit charges up C2 until the resistive potential divider (1M/120k) provides enough voltage to break down the diac D2. The triac TR1 then switches C3's charge into the pulse transformer, which ionises the tube into conduction using the charge on C2. The cycle then repeats. For slow/infrequent flash rates, small modifications to the circuit allow the converter section to shut down when the reservoir capacitor is charged, thus providing an efficiency enhancement.

Low Noise Audio Input Stage


+V
R3
Input
Zs = 1 to 2R (typ)
R1
R2
Output
NPN - ZTX650 or ZTX690B Series
PNP - ZTX750 or ZTX790A Series
Or I(sources)
-V

For low noise, low impedance amplifier applications such as moving coil input stages for hi-fi pre-amplifiers, and professional microphone buffer stages, it is essential to minimise noise generated by resistive components within the input loop. This is sometimes effected by using a parallel arrangement of devices or using large power transistors, both methods being valid ways in which the effective base resistance (rbb') seen by the circuit can be reduced. However, superior and reproducible results can be realised by employing the Zetex range of medium power devices.

This is due to the base resistance presented being much lower than with small signal devices. The absolute value of rbb' is difficult to define as it is extremely bias dependent, but is typically a few ohms. For comparative purposes, the noise factor produced by a ZTX650 transistor is 0.25dB compared with typically 2dB for a small signal type, at the industry standard test condition of 1 kHz, 5V, 200µA, and Rs = 2kΩ.

In the example circuit given, which is essentially a discrete component realisation of an op-amp, R1 effectively sets the input impedance, (cartridge manufacturers provide optimum load resistance values), while R3/R2 set the gain. Some optimisation may be achieved, dependant on application, by operating the input pair at seemingly high currents, (typically 10-40mA), thereby reducing the effects of emitter resistance. Current sources for the differential pair's tail current and the PNP's collector current could be replaced with current sources, similarly active loads for the diff' amp are also an option. Variations on the simple design shown are capable of yielding figures of better than -180dBv/Hz noise contribution per device.

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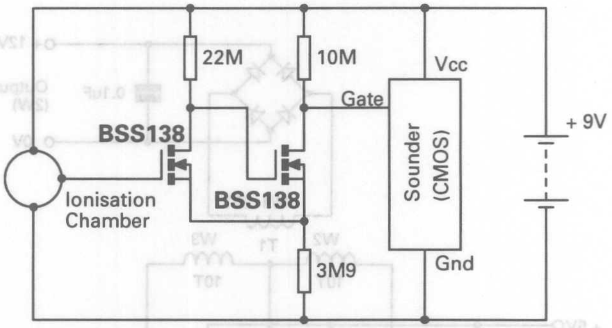
Ionising Smoke Detector



Smoke particles entering an ionising detection chamber generate signals of only a few pico-amperes. This signal must be buffered by a high input impedance amplifier, normally requiring an expensive low leakage J-FET.

The leakage current through the high integrity silicon-oxide gate of the Zetex BSS138 is negligible, hence the buffer circuit shown is ideal for this application. Two MOSFET's have been used, coupled to produce a Schmitt-Trigger action. This eliminates unwanted current drain produced by many CMOS IC's when their inputs are allowed to settle mid-supply. This circuit can also be used in signal conditioning circuits for battery operated CMOS microcontrollers. Suitable for operation over the temperature range of -40°C to +75°C the circuit consumes only 0.65µA (excluding the sounder), thus giving a battery life limited only by internal battery leakage.

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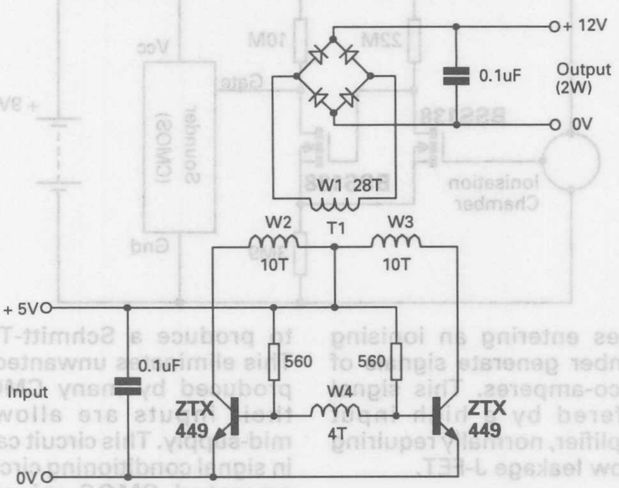
High Efficiency Hybrid DC-DC Converter



To meet the growing requirement of DC-DC converters that fit into 0.6 inch pitch DIL IC packages (for LAN hub power supplies for example), efficient but low cost hybrid circuits must be used. Using low switching loss bipolar transistors from Zetex, converters such as the circuit shown here can be constructed on small substrates giving power densities of over 1.7W/cc. Operating at over 80kHz the toroidal transformer and ceramic decoupling capacitor volumes are minimised, while the consistency of the h_{FE} process provides a balanced switching performance with no worry of low temperature start-up or transistor induced shoot-through.

The ZTX449 devices, possessing a saturation voltage of only 150mV at 0.5A give the 2W converter an efficiency of over 77%, thus keeping power losses and internal heating at a low level. Maximum operating temperature is ultimately limited by the magnetics but the standard industrial range of -20°C to +85°C is easily covered. Other devices are available that exhibit higher gain and/or higher voltage ratings for other supply voltage options.

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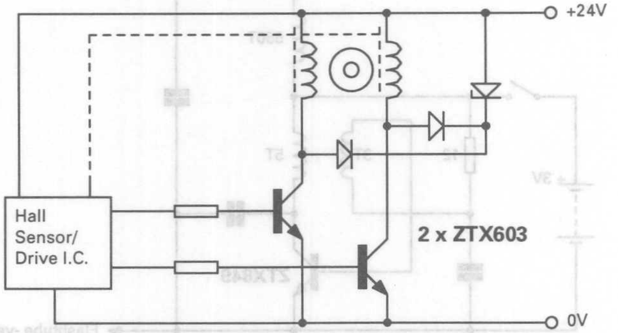
Brushless DC Fan Driver



High power brushless fans incorporate Hall effect rotor sensors which directly control stator winding driver transistors. Since these drivers must withstand high surge and continuous currents which occur in normal and stalled operation, expensive TO220 or TO126 Darlington transistors are commonly specified.

The 200°C and 1W capability of the E-Line (TO92 style) package coupled with the high efficiency matrix geometry of the ZTX603 Darlington transistor provides the necessary performance at a smaller size and cost. Featuring a saturation voltage of 0.95V at 1A for a base current of only 1mA, a 4A peak capability (1A continuous) and excellent switching characteristics, the ZTX603 is an efficient load tolerant driver for fans of powers up to 10W and above.

(Zetex manufacture a wide range of Darlington geometry transistors, with BV_{CEO} ratings from 30 to 140V, and DC current ratings upto 1A in TO92.)

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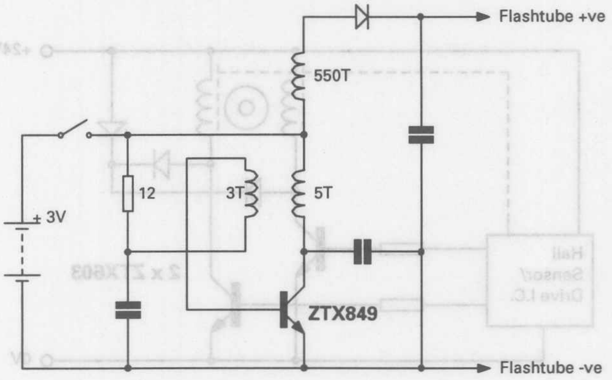
Flashgun Converter



Component size and capacitor re-charge times are a major consideration for camera flashguns. Regular TO220, TO126 or SM D-PAK packaged chips, although bigger, cannot match the performance of the E-Line (TO92 style) Matrix Bipolar chip family. Zetex transistors exhibit lower saturation voltages, higher h_{FE}, a higher current gain hold-up/silicon area, and a more efficient switching performance than any of the devices available from the aforementioned packages.

The ZTX849 features a saturation voltage of only 180mV at 5A (thus enabling a 5A continuous current capability in a TO-92 sized package), h_{FE} specified from 10mA and up to 20A, and an F_T of 100MHz thus allowing very efficient converter designs.

The simple circuit topology shown can provide a rapid capacitor recharge time of less than 5 seconds. Efficiency is also improved over competitive types, giving typically 24-36 more flash/recharge cycles per battery (Lithium).

A higher gain alternative, the ZTX869, is also available.

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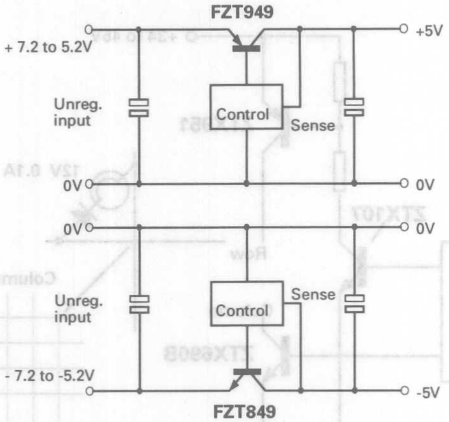
Low Drop Out Linear Regulators



Unlike standard regulator circuits and ICs which need at least 2V or more across the series pass transistor to operate, low drop out (LDO) regulators can function with only 100 - 200mV differential.

Low drop-out voltage regulators are essential for many battery powered applications including computers and other logic applications, in order to maximise useful battery life and reduce re-charge frequency.

Automotive electronics is another applicable area, where in engine management systems they are employed to ensure that the circuitry is correctly powered even when the automobile is cranking, and the battery's terminal voltage has fallen to 6 or 7V. They are also useful in mains powered applications where power dissipation / heatsink savings can be realised.

The ultra low saturation voltage of the ZBD849 and ZBD949 transistors enable these linear regulators to operate at currents of 5A continuously with drop-out voltages of only 180mV for the NPN circuit and 350mV for the PNP.

Packaged in the compact TO126, these high performance transistors are easy to mount and can be used free-standing for lower output current applications. Surface mount and through-hole options are available in the SOT223 and E-Line (TO92 style)packages.

The control circuit used can be a standard IC such as the LT1123 or a discrete design if the application requires extra flexibility.

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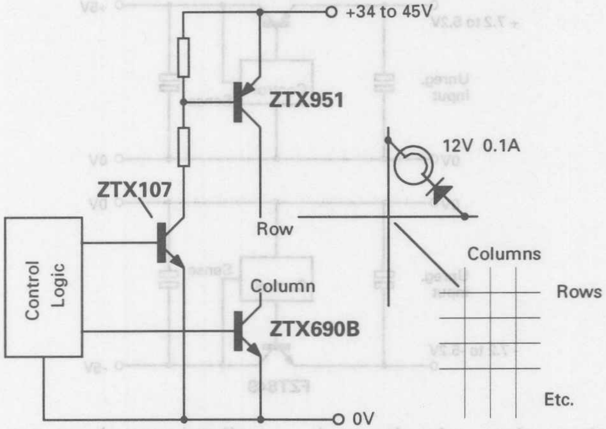
Gaming Machine Lamp Driver



x 8 for an 8 x 8 matrix of lamps

In slot machines and other applications where a large number of lamps must be controlled, matrix circuits are employed to radically reduce the number of lamp drivers required. Here, each row is selected in turn by control logic or more usually by an appropriately configured micro-controller. During row selection, the appropriate column drivers are activated to give the required display output. Consequently, each row driver must drive 8 lamps simultaneously, taking a worst case surge current of 6A. The column drivers must have very high gains to pass surges of up to 800mA using direct logic interfacing. TO126 or TO220 packaged Darlingtons are routinely selected for these drivers.

The ZTX951 and ZTX690B Super E-Line (TO92 style) transistors can easily replace these costly drivers. The high gain ZTX690B features a saturation voltage of 220mV at 800mA with just 4mA of base current allowing direct logic drive. The ZTX951 needs 100mA of base current to handle lamp turn-on surges, and will give a saturation voltage below 100mV in normal operation.

The number of bulbs driven can be easily increased by adding row drivers and increasing the supply voltage to maintain lamp power with reduced duty cycles. With a supply of 45V, 16 rows (128 lamps) can be accommodated.

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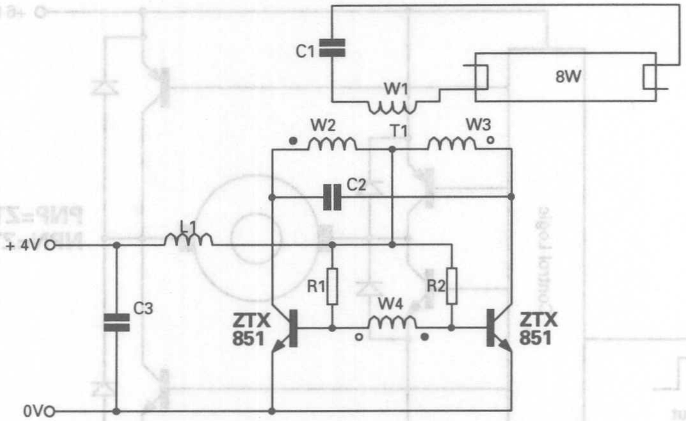
Emergency Lighting Fluorescent Lamp



The 8W emergency lighting converter shown here can be constructed at a fraction of the cost of older TO220 transistor based designs. Using the ZTX851, which has a saturation voltage of only 150mV at 4A (the cycle by cycle peak current seen with this design), the circuit runs with an efficiency of over 70%, a key feature for these battery operated circuits. Care with the transformer design and construction will allow the use of the higher gain ZTX869 which will raise efficiencies even higher. The ZTX851 and ZTX869 are available in the E-Line (TO92 style) package which has a 1.2W power rating for these device types. This permits significant space savings to be made since the bulky TO220 transistors and corresponding heatsinks can be eliminated. The operating frequency has been limited to around 25kHz to minimise transformer losses, yet ensuring the converter is inaudible.

Emergency lighting systems provide illumination in the event of mains failure and consist of a monitor circuit, a battery pack with trickle charging, and a DC-AC inverter. The inverter is used with the existing fluorescent tubes or an additional smaller tube, and is enabled by the monitor circuit.

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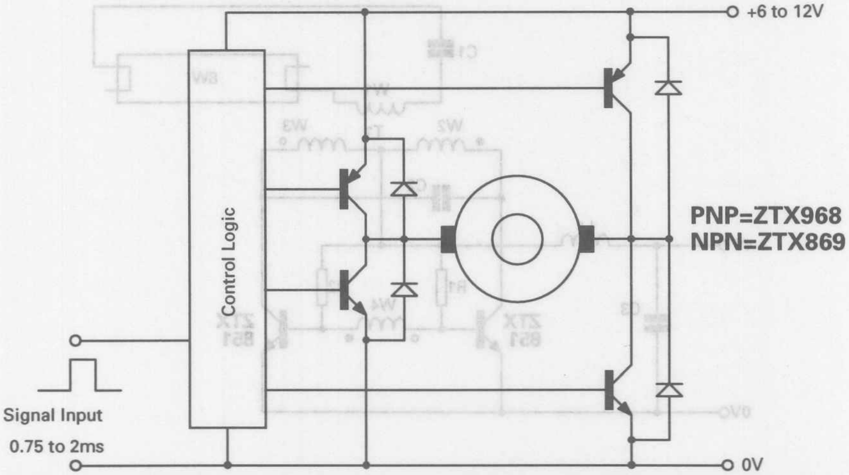
Remote Control Motor Driver



The combined features of low saturation voltage, a 5A continuous current capability, very high gain at high currents, and the compact dimensions of the TO92 style E-Line package, provide an attractive alternative to TO220 devices in this commonly used driver circuit.

The ZTX869 and ZTX968 produce typical $V_{CE(sat)}$ values at 5A collector current, of 180mV and 250mV respectively, and possess useful gains of 100 and 50 at 20A (typical: for the '869 and '968), ensuring cool and efficient operation under surge conditions. It is also possible in some designs to omit the collector-emitter diodes, as the very high inverse gain inherent to the Zetex process (which in low voltage variants can be approximately 33 to 50% of the forward h_{FE}) acts to conduct inductive transients. This prevents the subsequent degradation to the emitter-base junction, as must be considered and accounted for in conventional designs.

These features make it possible to construct this control unit, for electric motors with input power requirements to over 40W continuous, at a fraction of the size and yielding a typical 4 times cost saving over the present MOSFET based designs.

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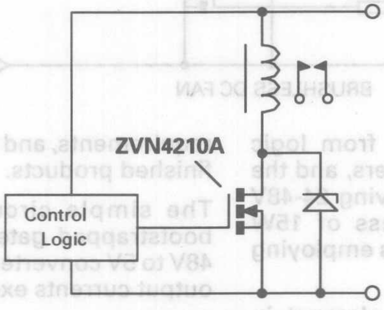
Features and Applications of the ZVN4210A

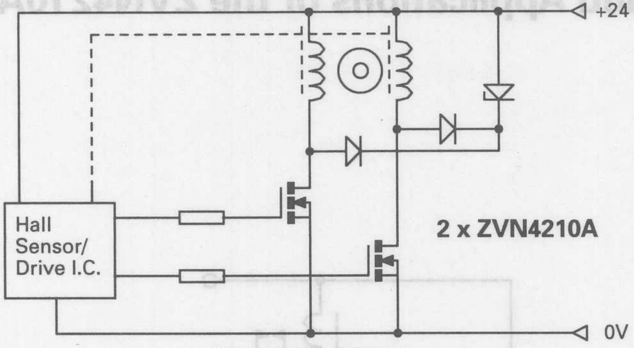


RELAY DRIVER

The ZVN4210A is one of the latest E-Line packaged developments available from Zetex. It is a cellular design, N-Channel Enhancement mode MOSFET that features a 100V drain-source voltage rating, an on-resistance of 1.5Ω maximum (1.1Ω typical) at 1.5A, a current rating of 0.45A continuous and up to 6A under pulsed conditions. These attributes, in addition to the usual MOSFET advantages of fast switching, ease of drive requirements, and freedom from secondary breakdown, make the ZVN4210A an excellent choice as a logic to power interface. This enables reliable minimum part count designs to be realised at low cost.

The low switching losses produced by the fast switching speeds, (T_{on}=8ns; T_{off}=28ns typical, for a drain current of 1.5A and a V_{gs} of 10v), ensure cool and efficient operation in switching circuits.

The ZVN4210A should prove to be of interest to the Automotive Electronics sector, where it's inherent ruggedness, non-critical drive requirements, and low losses provide an attractive alternative to the TO220 packaged Darlingtons often employed as relay drivers. Of particular concern to designers are the requirements for surviving fault conditions. The ZVN4210A greatly simplifies the problem as no bias current/gain considerations are necessary, due to it being a voltage driven device.



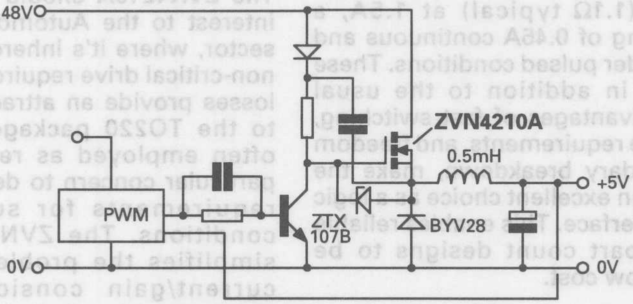
BRUSHLESS DC FAN

Brushless DC fans benefit from logic driven, load insensitive drivers, and the ZVN4210A is capable of driving 24-48V fans with powers in excess of 15W (input), and higher in circuits employing locked rotor protection.

As the primary switching element in DC-DC converters, Zetex MOSFETS allow greater power density designs to be realised. Switching frequencies extending into Mega-Hertz are possible, lowering inductance and smoothing requirements, and thereby the size of the finished products.

The simple circuit shown features bootstrapped gatedrive, to enable this 48V to 5V converter to operate reliably at output currents exceeding 1.7A.

Other typical applications include lamp, solenoid, and motor drivers, and analog applications such as audio driver stages and high power analog switching networks.



48V to 5V 1.7A DC-DC CONVERTOR

ZETEX

Design Note 21
Issue 2 October 1995

Automotive Alarm Siren Driver

2A DC Rated Super-ß Transistors Provide Reduced Component Count 'H' Bridge

Introduction

The majority of modern automotive alarm sirens employ a moving coil loudspeaker, that is driven by an 'H'-bridge output stage. This output driver circuit is commonly effected with TO-126/TO-220 packaged power transistors, as the currents involved and the resulting power dissipation are both in excess of common smaller package products. Zetex Super E-Line (TO-92 style) transistors however, due to their superior device geometries and silicone packaging, can, and have been used in this application to provide a minimal drive circuitry topology, and a size, cost and performance advantage.

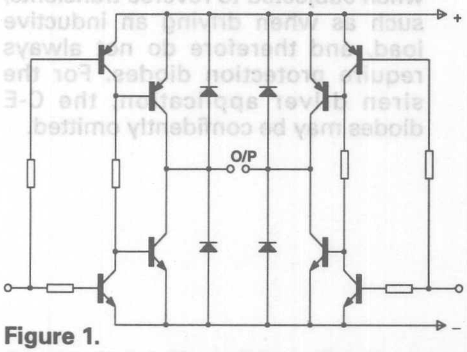


Figure 1.
Conventional Siren Driver Circuit using Separate Pre-driver Transistors.

Background

The circuit diagrams presented in Figures 1 and 2, show two common methods of effecting a siren driver circuit. These circuits include collector-emitter (C-E) diodes to protect the transistor in the event of negative rail excursions, due to inductive load transients, and base-emitter (B-E) resistors to provide a path for any transistor leakage current. This latter point is an important consideration when using the usual TO-126/TO-220 power products, as their high $V_{CE(sat)}$ can lead to appreciable temperature rises.

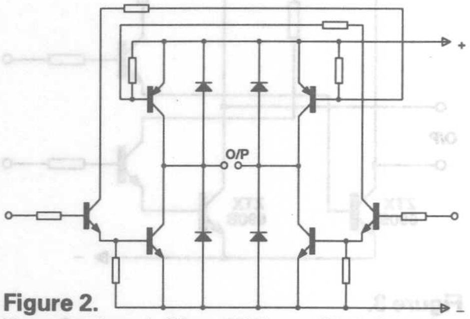


Figure 2.
Conventional Siren Driver Circuit using Cross-coupled Darlington Pre-driver Configuration.

Zetex E-Line transistors are optimised to possess a very low $V_{CE(sat)}$, and are therefore much more efficient in high current switching applications than conventional power products.

Some siren manufacturers use TO-220 packaged Darlington power transistors. While this can reduce the component count, (though can also increase the component cost), as the C-E diodes are sometimes integral to the Darlington, the intrinsic high on-state voltage of the Darlington configuration increases device dissipation. Consequently the circuit requires B-E resistors, and also reduces the power delivered to the load.

Circuit description

The circuit shown in Figure 3 has been devised using Zetex Super E-Line transistors. It is basically a modified version of figure 2. There are two important differences to note:

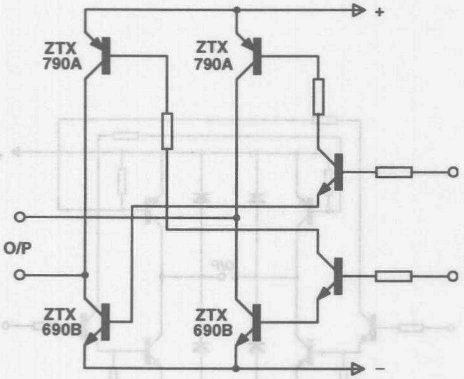
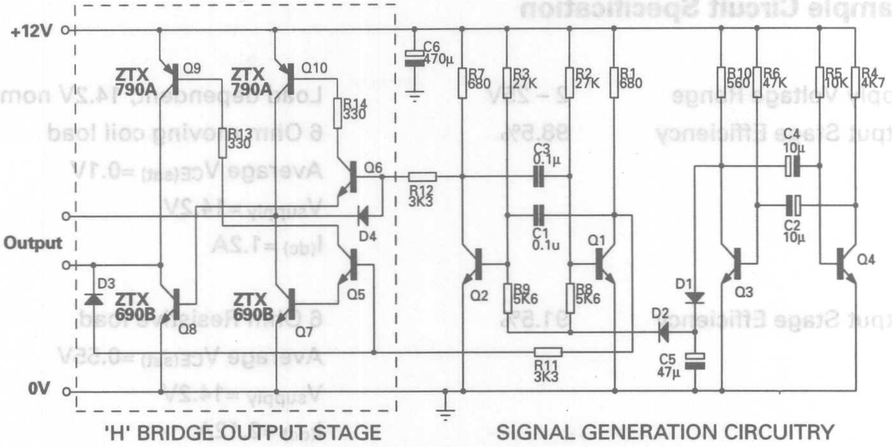


Figure 3.
ZETEX Reduced Component Count 'H'-Bridge Circuit.

i) As the devices employ an efficient matrix die geometry, they produce low $V_{CE(sat)}$ values, this leads to very low device dissipation, and therefore low temperature rise. (Viz. ZTX690B: $V_{CE(sat)} = 0.3V$ at I_C=2A, I_B =20mA and ZTX790A: $V_{CE(sat)} = 0.4V$ at I_C=2A, I_B =50mA, please consult ZTX688B-ZTX696B and ZTX788B - ZTX796A series datasheets). This can be very important in automotive applications, where the operating environment can reach high ambient temperatures. It follows that in this case, the usual base-emitter resistors employed to aid turn-off, and to divert leakage currents at high junction temperature, are no longer required.

ii) The Zetex devices chosen, possess a reasonable h_{FE}, (strictly termed h_{FC}) when used in the inverted mode, Eg. with the collector/emitter terminals reversed. Typical peak values being 125 for the ZTX690B and 100 for the ZTX790A . This feature means that they are still capable of operating when subjected to reverse transients, such as when driving an inductive load, and therefore do not always require protection diodes. For the siren driver application, the C-E diodes may be confidently omitted.



Q_1 - Q_6 MPS2222A
Q_7 - Q_8 ZTX690B
Q_9 - Q_{10} ZTX790A
D_1 - D_4 1N4148

Figure 4 .
Example Circuit (Note 1).

The modified 'H'- bridge siren driver stage has been incorporated into a complete siren driver circuit, in order to demonstrate the capability of Zetex Super E-Line transistors. The demonstration circuit is shown in figure 4, and consists of two sections; the signal generation circuitry, and the 'H' bridge output stage. The former comprises a slow running astable multivibrator to generate a ramp signal, that is then used to modulate a second multivibrator, causing it to sweep over a range of frequencies. This section drives the 'H'-bridge output stage described previously.

While the multivibrator method is valid, (some manufacturers use a similar circuit, while others favour op-amp derived circuits, or even small micro-controllers - see Note 2), there can be a problem with cross-conduction, Eg. Simultaneous conduction by both NPN and PNP devices during switch-over, thus leading to excessive power dissipation. To prevent this, the output stage on the demonstration circuit has two additional signal diodes, D3 and D4. These components inhibit the driver transistors Q5 and Q6, until the opposing half of the 'H'-bridge has started to switch off. More elaborate signal generators incorporate delays between changes of output state, (using an extra gate or software) so would not require these extra diodes.

The circuit has been designed assuming a load impedance of 6 Ohms; a common value for the drive units used. Lower loads can be driven but may require either a change to the bias resistors, R13/14, to provide sufficient base current, or for very low impedance loads, a change of output transistors:- the ZTX851/ZTX951 with a continuous current rating of 5 Amps and 4 Amps respectively are eminently suitable.

Example Circuit Specification

Supply Voltage Range	2 – 25V	Load dependent; 14.2V nominal
Output Stage Efficiency	98.5%	6 Ohm moving coil load Average $V_{CE(sat)}$ =0.1V V_{supply} =14.2V $I_{(dc)}$ =1.2A
Output Stage Efficiency	91.5%	6 Ohm Resistive load Average $V_{CE(sat)}$ =0.55V V_{supply} =14.2V $I_{(dc)}$ =2.12A
Junction Temperature Rise	19°C	6 Ohm moving coil load V_{supply} =14.2V
Control ramp	1 to 9V, 4Hz	V_{supply} =14.2V
Swept frequency range	870Hz – 1.9kHz	V_{supply} =14.2V

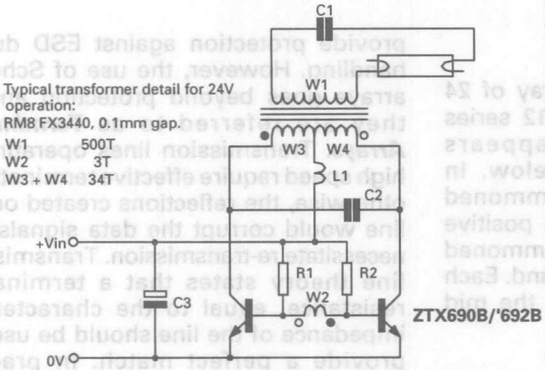
Note 1:
D3 and D4 are not required for signal generation circuits possessing output/output deadtime.

Note 2:
ZETEX now offer an IC solution for the signal generation circuitry - namely the ZSD100.

ZETEX

Design Note 22
Issue 2 June 1995

LCD Display Fluorescent Backlighting



Typical transformer detail for 24V operation:
RM8 FX3440, 0.1mm gap.

W1	500T
W2	3T
W3 + W4	34T

Note 1: For this circuit topology the collector-emitter only experiences a high voltage when the base has been driven negative by the feedback winding. This means that the voltage rating most relevant is the BV_{CEV} or BV_{CES} parameter.

Consideration of this feature will allow selection of a lower rated BV_{CEO} device for the application, which in turn, will provide a lower $V_{CE(sat)}$ and hence higher conversion efficiency.

LCD displays are becoming more common in such diverse areas as laptop computing and supermarket till systems. As the LCD display itself does not generate light, a small fluorescent tube and diffusing panel are usually integrated into the unit. The tube is driven by an invertor that must (typically) generate 1000V ptp to initiate conduction, and 500V ptp while the tube is running. The unit must be compact, efficient, and probably be operable over a wide voltage range, to enable contrast adjustments.

The push-pull invertor shown, running at an inaudible 30kHz was designed as a replacement option for a commercially available 24V unit. It uses Matrix geometry Super E-Line (TO92 style) transistors that offer a more efficient performance than the "tall TO92" devices sometimes utilised; a compact RM series transformer core; and a few other components to allow this design to be constructed at a fraction (estimated to be 50%) of the cost of pre-built units.

The ZTX690B and ZTX692B devices are part of the ZTX688B-696B high gain series; the transistor choice being dependant on the operating voltage range. These particular devices possess BV_{CEO} ratings of 45V and 70V, (see note 1 below) yet still provide typical $V_{CE(sat)}$ values at 500mA of 50mV and 80mV respectively. For lower power requirements, the high performance FMMT489 and FMMT491 SOT23 transistors should also prove attractive. (For optimum performance the "SuperSOT" FMMT618 series should be considered).

[Also refer to AN14]

Features and Applications of the SDA24 Schottky Array

Bus Termination for High Speed Data Systems

SDA24 - Schottky array

The SDA24 is a monolithic array of 24 Schottky diodes, arranged as 12 series pairs. The diode array appears schematically in Figure 1 below. In practice the cathodes are commoned together and connected to the positive rail, (Vcc) in the system; the commoned anodes being connected to ground. Each data line is then connected to the mid point of each series pair.

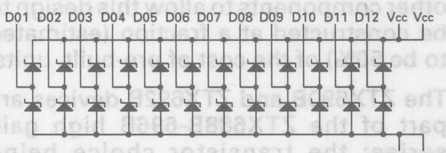


Figure 1
Schottky Barrier Diode Bus Termination Array

Transients occurring on the data or clock lines of memory or LSI Logic devices (e.g. ***DRAM, SRAM***, Hard disc controllers), can result in improper operation, data loss or even damage to the devices' input circuitry. These transients can be caused by transmission line reflections, crosstalk between lines, switching noise and external influences such as radiated supply glitches.

The first devices to use a diode clamp were CMOS logic gates, in order to provide protection against ESD during handling. However, the use of Schottky arrays goes beyond protection and so they are referred to as ***Termination Arrays***. Transmission lines operating at high speed require effective terminations, otherwise, the reflections created on the line would corrupt the data signals and necessitate re-transmission. Transmission line theory states that a terminating resistance, equal to the characteristic impedance of the line should be used to provide a perfect match. In practice however, use of resistive termination networks in digital systems can lead to an increased power consumption, inferior noise immunity and, for some networks increased propagation delays.

Schottky diode arrays do not suffer from any of the above drawbacks. Under normal conditions, the diode pairs connected to each line are almost transparent to the data signal - the only effects being a slight capacitive loading due to the depletion capacitance, and any diode reverse leakage current. However, if the voltage at the junction of the series pair increases above V_{CC} or decreases below ground, due to a transient or pulse reflection, then the upper (or lower) diode respectively will start to conduct. This dissipates the over/under voltage to the appropriate supply rail. The series pair can then be seen as a voltage clamp that is effective at a Vf above and below the

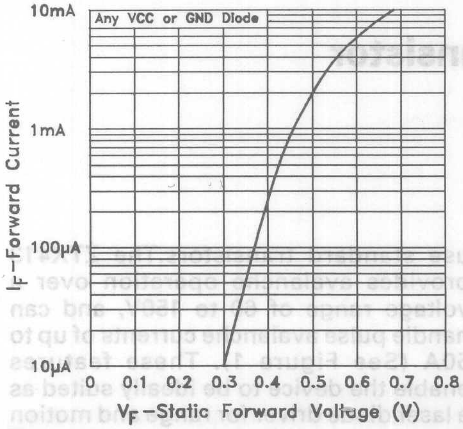


Figure 2
Low I_F vs V_F Characteristic

supply rail. E.g. on a 5V supply, each Schottky pair will contain the data signal within a 5V + Vf and 0V-Vf window, where the Vf will be typically 0.4V at low currents. (Please refer to figure 2).

The waveforms in Figure 3 illustrate this action, and were produced by driving a 10m length of typical digital comms coaxial cable. The effect of voltage clamping can be seen on the first overshoot, and the subsequent reflections are reduced almost to zero.

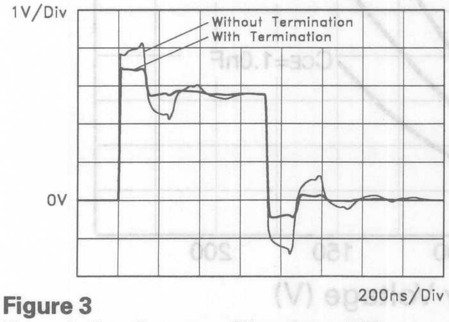


Figure 3
Example Scope Display Effect of terminating Schottky pair on a pulse transmission

Applications

Applications for these arrays can be found in areas where high speed parallel data communication is required such as:

- Termination networks
- Network peripherals
- Hard disc drives
- CD ROM drives
- Floptical Drives
- DAT drives
- Disc arrays

Glossary

RAM-Random Access Memory - an LSI device used to store binary data. The data is stored within a matrix of cells that are referenced using an address. The device is able to be written to/read from in any order, (hence random), and is a "volatile" store - if the supply is removed, the data will be lost.

DRAM-Dynamic RAM - this form of RAM needs to be continually refreshed, otherwise the data will be lost, and so requires an additional refresh controller, though some microprocessors supply this function automatically.

SRAM-Static RAM - this doesn't require a refresh function as does dynamic RAM. It will store data written to it until it is changed by the system or lost by a power loss or shut down.

CROSSTALK-The noise or signal leaked or induced from one circuit, bus or IC section to another.

Design Note 24
Issue 1 October 1995

The ZTX413 Avalanche Transistor

Low Voltage Operation up to 50A

The ZTX413 is the latest addition to the Zetex range of avalanche mode transistors. The ZTX413, in common with the ZTX415 through-hole device, and FMMT415 and FMMT417 surface mount components, are processed and characterised specifically for avalanche mode operation. They can thereby offer advantages such as reliability and reproducible avalanche parameters and circuit performance; factors very difficult to control when attempting to use standard transistors.The ZTX413 provides avalanche operation over a voltage range of 60 to 150V, and can handle pulse avalanche currents of up to 50A (See Figure 1). These features enable the device to be ideally suited as a laser diode driver for range and motion measurements. Simple charge storage methods can be utilised; the circuit shown (see Figure 2) produces a 20A sinusoidal-like pulse of 20ns base width using a 100V supply.

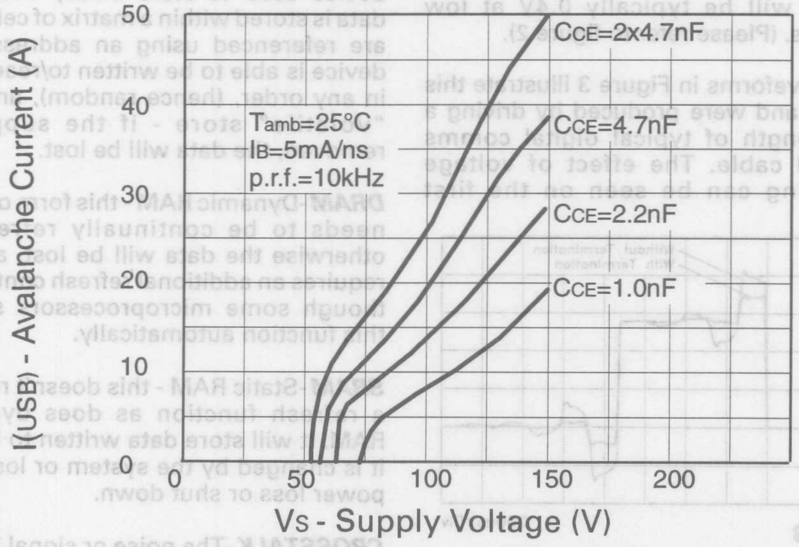


Figure 1.
Avalanche Current against Supply Voltage.

The avalanche current is dependent on the supply voltage, the collector-emitter capacitance, and load and PCB layout inductance. The chart shown indicates the typical pulse current produced for C-E capacitance values of 1nF to 9.4nF (actually 2 off 4n7), over the supply voltage range. The capacitor used for this application must be a low loss, low inductance type. Ceramic chip capacitors (as used for RF circuits) and small, thick leaded disc parts were used for the characterisation of avalanche parameters. The circuit boards are best constructed using a ground-plane technique, and all component leads in the discharge loop must be as short as possible, and be of very low resistance.

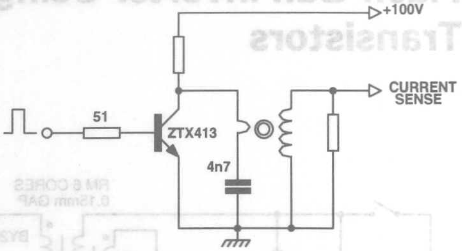


Figure 2
Avalanche Transistor Test Circuit.

Appendix

Papers Detailing Applications Of Avalanche Transistors

Application Of Avalanche Transistors To Circuits With A Long Mean Time To Failure
Werner B.Herden.
IEEE Transactions On Instrumentation And Measurement, Vol.25, No.2 June 1976.

Properties Of Avalanche Injection And Its Application To Fast Pulse Generation And Switching
Yoshihiko Mizushima and Yoshiharu Okamoto.
IEEE Transactions On Electron Devices, Vol. ED-14, No. 3, March 1967

Static And Dynamic Behaviour Of Transistors In The Avalanche Region
Paolo Spirito
Istituto Elettrotecnico, Universita di Napoli, Piazzale Tecchio, 80125 Napoli, Italy
IEEE Journal Of Solid-State Circuits, April 1971.

An Analysis Of The Dynamic Behaviour Of Switching Circuits Using Avalanche Transistors
P. Spirito and G.F. Vitale
IEEE Journal Of Solid-State Circuits, August 1972.

Flash Gun Inverter Using Super E-Line Bipolar Transistors

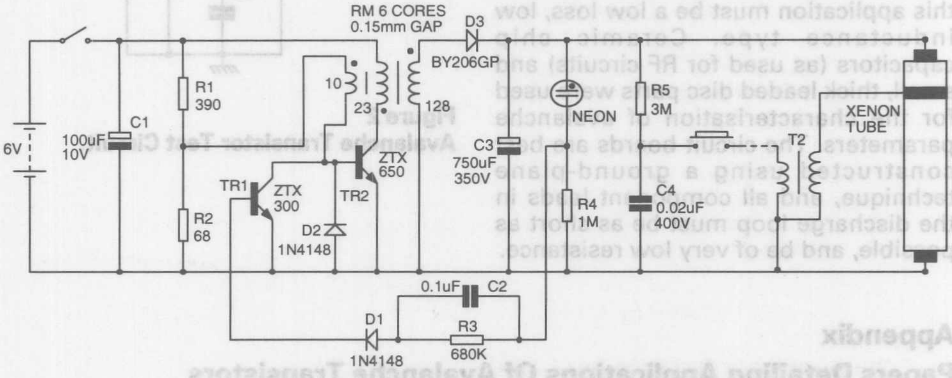


Figure 1.
Flashgun Circuit using Flyback Inverter.

The Xenon Flash tube considered for this design requires 250V to supply the flash energy of 20 Joules maximum, and about 4kV to ionise the gas to trigger it into conduction. The energy source available is a 6V battery. A flyback inverter is utilised to provide DC-AC inversion to charge the main reservoir capacitor. A suitable transistor for this circuit must possess a high enough voltage rating to withstand flyback transients, yet be capable of conducting sufficient current to pass a relatively large amounts of energy to the inductor. This combination of parameters would normally dictate that a large device would be necessary. However, the ZTX650 E-Line transistor - a 60V, 2A rated TO92 style product, allows an efficient, cost and size optimised design.

The operation of the circuit is as follows:

On switch-on, the transistor Tr2 is biased on by the resistors R1, R2 and a small voltage is generated across the primary of the transformer. The transformer action induces a voltage in the feedback winding which turns the transistor hard-on. Whilst the transistor is on, the energy is stored in the primary inductance of the transformer, and is delivered to the output circuit when the transistor turns off. That is, the stored magnetic energy is converted into electrical energy to charge the output capacitor. When the transistor is hard-on or saturated, practically the whole of the supply voltage is applied across the primary winding. Thus, linearly increasing collector current flows through the transistor, whilst the controlling base current I_B flows into the base. The transistor remains saturated until the collector current exceeds a value $\beta.I_B$ (β is the static common emitter current gain of the transistor). The transistor begins to come out of saturation and the collector voltage rises. The voltage across the primary thus decreases which also encourages a decrease in base current. This action continues until the voltages across the windings are reversed cutting off the transistor. The stored energy is now dumped into the output capacitor and secondary current decays to zero. For every cycle, an energy equal to $0.5Lp.Ipk^2$ is dumped into the capacitor (Lp - the primary inductance, Ipk - the peak current flowing in the primary prior to switching off).

With the component values shown in the diagram, the unit takes about 7 seconds to charge up fully, and will supply a flash energy of 20 joules.

When the trigger button is pressed, the capacitor C4 is discharged via the primary of the pulse transformer which generates a pulse of about 4000 volts on the secondary. This pulse is applied to the trigger electrode of the Xenon tube causing it to strike. The flash tube then discharges capacitor C3, giving a very intense flash of light. The intensity will depend on the value of C3 and the voltage to which it is charged.

To keep power consumption down, the circuit is designed so that when the output capacitor has charged up to the required voltage, the inverter switches itself off. The operation is as follows:

Initially, the voltage across C3 will be below the final value. The neon L1 will not be struck thus the voltage across R4 will be zero, Tr1 will not be conducting and the inverter operates normally. When the voltage across C3 reaches the final value, the neon L1 will strike causing a voltage drop across R4. This biases Tr1 on which keeps the transistor Tr2 off, thus inhibiting the inverter action. The charge on C3 then slowly decreases due to normal leakage and the small amount of current supplied to the neon and Tr1. When the voltage across C3 reduces to about 200 volts, the neon extinguishes, thereby turning Tr1 off and the inverter starts to re-charge C3.

The capacitor C2 is used as a speed-up capacitor. D1 and D2 protects the base-emitter junction of Tr1 and Tr2 respectively.

Design Note 26
Issue 2 January 1996

D.C. Motor Speed Control using the ZTX450

Introduction

The function of the following circuit is to improve the load/speed regulation of a d.c. machine. One of the main reasons why the speed of a permanent magnet field d.c. motor varies with load, is that a voltage drop is current and hence load dependent. The circuit given in Figure 1 stabilises the speed of the motor by cancelling out the effect of the motor rotor resistance using a bridge circuit.

Theory

If the bridge and variable resistor are analysed on their own (Figure 2), a proof can be derived which shows that the control voltage (i.e. the speed) is not directly dependent on the motor current. (Please refer to Appendix).

The value of r_a varies from motor to motor, so the bridge must be balanced to suit the motor employed.

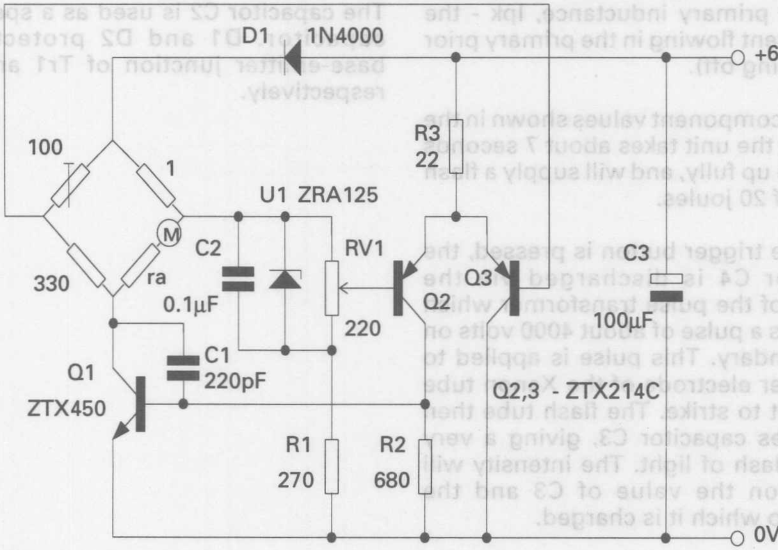


Figure 1
DC Motor Speed Control Circuit.

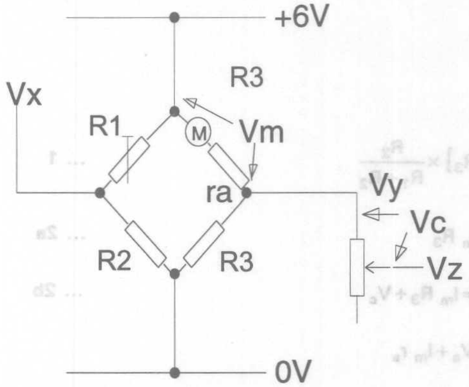


Figure 2
Bridge Circuit.

This can be done by calculation and measurement of the motor resistance, or by adjusting the value of the 100Ω preset resistor until the motor speed just becomes unstable and then backing off a fraction.

The current/speed and torque/speed characteristics of a test motor have been plotted in Figures 3 and 4, with and without the control circuit, to show the improvement in speed stability. Variations in motor speed due to supply changes are also greatly reduced by the circuit. On the test motor a speed change of ± 2% was recorded for a supply change of ± 20%.

Larger motors can be accomodated by using higher current capable transistors, such as the ZTX689B - a high gain, 2A DC rated device, or the ZTX869 - a high gain 5A DC rated transistor.

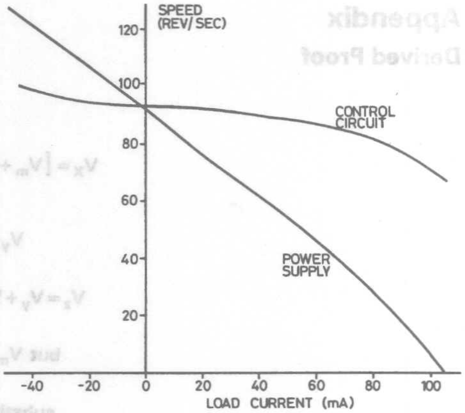


Figure 3
Current/Speed Characteristics of a Test Motor.

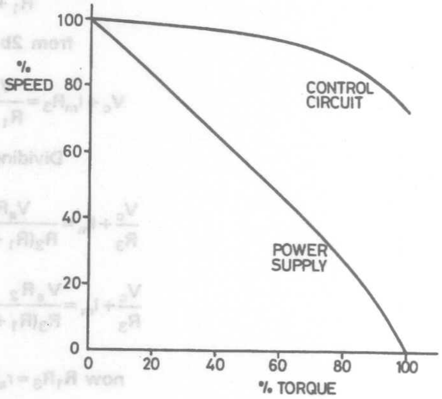


Figure 4
Torque/speed Characteristics of a Test Motor.

Appendix

Derived Proof

$$V_X = \left[V_m + I_m R_3\right] \times \frac{R_2}{R_1 + R_2} \qquad \ldots\ 1$$

$$V_y = I_m\ R_3 \qquad \ldots\ 2a$$

$$V_z = V_y + V_c = I_m\ R_3 + V_c \qquad \ldots\ 2b$$

but $V_m = V_a + I_m\ r_a$

substituting in 1.

$$V_x = \left[V_a + I_m r_a + I_m R_3\right] \times \frac{R_2}{R_1 + R_2} = \left[V_a + I_m(r_a + R_3)\right] \times \frac{R_2}{R_1 + R_2} = \frac{V_a R_2}{R_1 + R_2} + I_m R_2 \frac{(r_a + R_3)}{R_1 + R_2}$$

now $V_x = V_z$ (1st assumption)

$$\text{so } V_z = \frac{V_a R_2}{R_1 + R_2} + I_m R_2 \frac{(r_a + R_3)}{R_1 + R_2}$$

from 2b $V_z = I_m R_3 + V_c$

$$V_c + I_m R_3 = \frac{V_a R_2}{R_1 + R_2} + I_m R_2 \times \frac{r_a + R_3}{R_1 + R_2}$$

Dividing by R_3 gives

$$\frac{V_c}{R_3} + I_m = \frac{V_a R_2}{R_3(R_1 + R_2)} + I_m R_2 \frac{r_a + R_3}{R_3(R_1 + R_2)}$$

$$\frac{V_c}{R_3} + I_m = \frac{V_a R_2}{R_3(R_1 + R_2)} + I_m \left[\frac{R_2 r_a + R_2 R_3}{R_1 R_3 + R_2 R_3}\right]$$

now $R_1 R_3 = r_a R_2$ (2nd assumption)

$$\frac{V_c}{R_3} + I_m = \frac{V_a R_2}{R_3(R_1 + R_2)} + I_m$$

$$V_c = \frac{V_a R_2}{R_1 + R_2}$$

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Design Note 27
Issue 1 February 1996

12V Solenoid Driver/Latch

Single Switch Operated, High Current High Side Switch

This circuit has been designed to control a solenoid by the operation of a single push-button switch. It will supply loads of over 1A and can be operated up to a maximum speed of once every 0.6 seconds. When power is first applied to the circuit, the solenoid will always start in it's off position. Other features of the circuit are it's automatic turn-off if the load is shorted, and virtually zero power consumption when off.

When the supply is connected, the 470Ω and 270Ω base-emitter resistors ensure all three transistors remain off. The 1µF capacitor charges up to a value approaching that of the supply rail. If the push-button switch is then closed, the charge is transferred to the bias network of the ZTX300, turning it on. This, through the 100Ω and 270Ω bias resistors of the ZTX550s turn these devices on which energises the load. This action also holds the ZTX300 on (via the two 560Ω resistors), irrespective of the decaying charge on C1.

Since the ZTX300 is now on, when the push-button is released the 1µF capacitor will be discharged through the transistor via the bleed resistor.

If the push-button is operated again it will connect this discharged capacitor to the bias network of the ZTX300 turning it and thus the output off. Any excess energy stored in the solenoid will be dissipated in the 1N4000 protection diode.

When the push-button switch is released, the 1µF capacitor will charge up ready to trigger the latch on again when the switch is operated. The 0.1µF capacitor inhibits false triggering due to transient voltages.

This circuit was originally designed for an Automotive application, but similar circuits should also find use in industrial control systems. Similar circuits for higher current applications are possible with a change to the output device. For example, the ZTX751 2A rated PNP, or the high gain ZTX790A could be used for load currents of up to 2A, while the ZTX951 allows operation to over 3A.

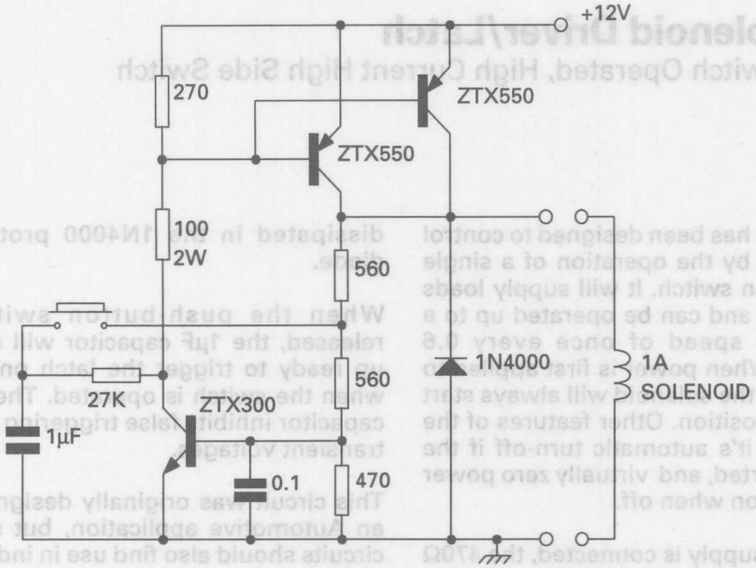


Figure 1
12V Latch Circuit.

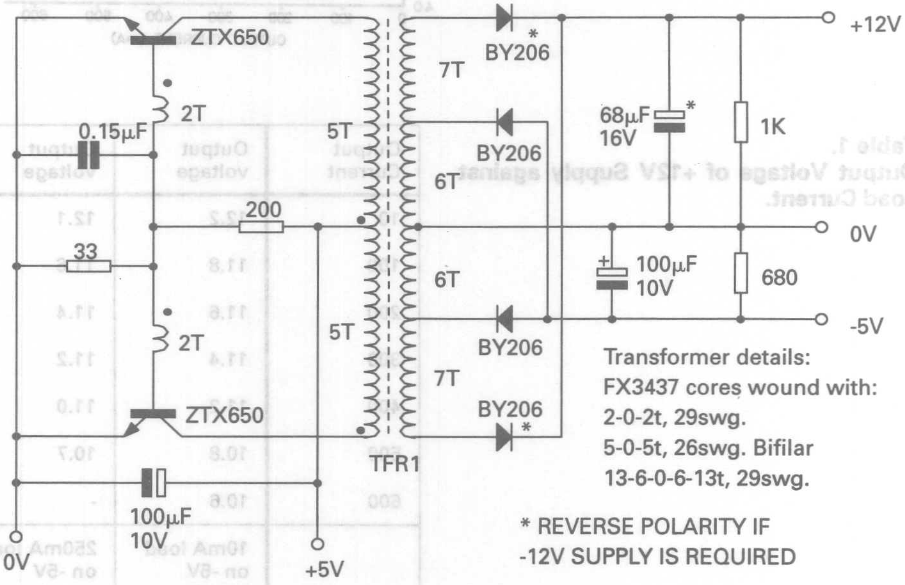
ZETEX

Design Note 28
Issue 2 January 1996

6W Inverter for MOS Logic supplies

The 6W inverter shown in Figure 1 has been designed to generate the extra power supplies required by popular MOS memories from a normal 5V TTL supply source. It may be used to supply up to eight 2808 read only memories which require supplies of +5V, -5V and +12V, or if the output components of the 12V section are reversed, the circuit will power over ten 5204 ROMs which require +5V and -12V supplies.

The inverter is a simple push-pull circuit which takes advantage of the high current handling capability of the ZTX650 range. It oscillates at a frequency of approximately 25kHz to allow it to use a very small transformer (RM6), and also to render the inverter inaudible. The output characteristics are given in Figures 2 and 3. Output ripple is approximately 150mV peak to peak on both outputs.



Transformer details:
FX3437 cores wound with:
2-0-2t, 29swg.
5-0-5t, 26swg. Bifilar
13-6-0-6-13t, 29swg.

* REVERSE POLARITY IF
-12V SUPPLY IS REQUIRED

Figure 1
6W Inverter Schematic.

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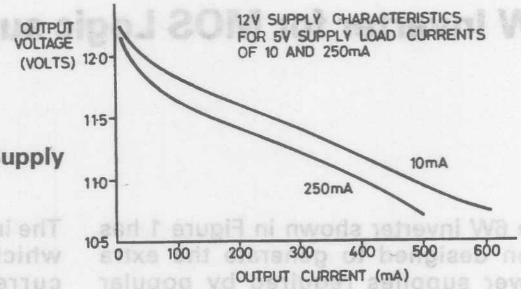


Figure 2
Output characteristics of +12V Supply against Load Current.

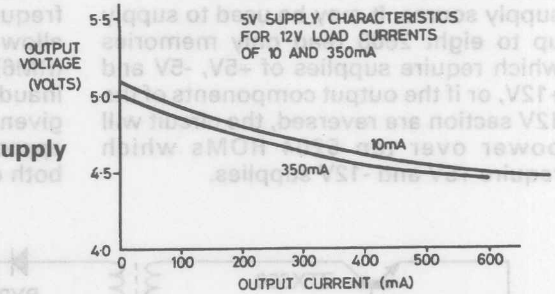


Figure 3
Output characteristics of -5V Supply against Load Current.

Table 1.
Output Voltage of +12V Supply against Load Current.

Output Current	Output voltage	Output voltage
10	12.2	12.1
100	11.8	11.6
200	11.6	11.4
300	11.4	11.2
400	11.2	11.0
500	10.8	10.7
600	10.6	-
	10mA load on -5V supply.	250mA load on -5V supply.

Design Note 29
Issue 1 June 1996

ZVP4424 and ZVN4424 in Telephone Hook-Switch and Earth Recall Applications

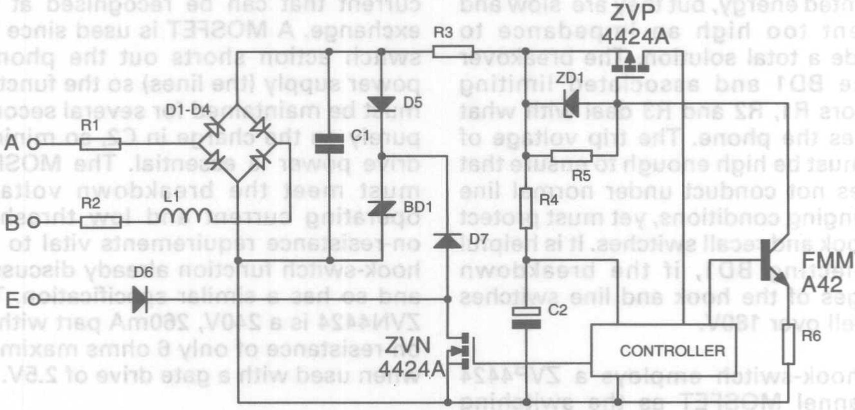


Figure 1
Line Interface Circuitry

Figure 1 shows the line interface circuitry of a "feature" telephone. Its major components are:- An input bridge rectifier, a transient protection circuit, an electronic hook switch, an earth recall switch and a logic controller.

The input bridge rectifier is necessary as the polarity of the voltage input lines is not defined (and for signalling purposes, can be switched). The line input is a DC supply of around 50V with a source resistance that can be as low as 500 ohms. When the phone is taken off-hook, it clamps the line input at a low voltage, causing a line current that is recognised by the exchange and captures the line. The off-hook current can vary from 30mA to 120mA depending on line length (resistance). When on-hook, the phone can take a small current from the line to charge batteries, maintain memories etc., but this current must be kept low if unwanted line capture is to be avoided. The exchange initiates ringing by adding a high voltage low frequency signal to the line. This boosts the normal line input voltage of the phone to around 180V peak.

The maximum input voltage to the phone can be much higher however. Lightning strikes occurring near telephone lines can induce voltage transients of several thousand volts sourced from a low impedance. Gas discharge suppressors mounted in dwelling entry sockets kill much of this unwanted energy, but they are slow and present too high an impedance to provide a total solution. The breakover device BD1 and associated limiting resistors R1, R2 and R3 deal with what reaches the phone. The trip voltage of BD1 must be high enough to ensure that it does not conduct under normal line and ringing conditions, yet must protect the hook and recall switches. It is helpful in selecting BD1, if the breakdown voltages of the hook and line switches are well over 180V.

The hook-switch employs a ZVP4424 P-Channel MOSFET as the switching element. With a BV_{DSS} of 240V, it makes selection of BD1 relatively easy. Since when the hook-switch turns on, it pulls the line voltage down to a few volts, it is vital that the switch can give a low on-resistance with low gate drive. The ZVP4424 has an on-resistance of only 11 ohms, even with just 3.5V gate drive. This low resistance allows the values of R1, R2 and R3 to be maximised (improving transient resistance) without raising the total phone resistance so high that it compromises long line performance. Since the MOSFET has a continuous current rating of 200mA, it can handle short lines well too.

Although the polarity of the phone input lines is not fixed, one line is always negative with respect to ground. This allows the Earth Recall signalling function to be achieved. The ZVN4424 N-Channel MOSFET shown in Figure 1 grounds whichever input line is most negative, causing a change in line current that can be recognised at the exchange. A MOSFET is used since the switch action shorts out the phone's power supply (the lines) so the function must be maintained for several seconds purely on the charge in C2, so minimal drive power is essential. The MOSFET must meet the breakdown voltage, operating current and low threshold on-resistance requirements vital to the hook-switch function already discussed and so has a similar specification. The ZVN4424 is a 240V, 260mA part with an on-resistance of only 6 ohms maximum when used with a gate drive of 2.5V.

(Both the ZVN4424 and the ZVP4424 MOSFETs are available in through-hole (E-Line, TO92 style) and surface mount (SOT223) packages. The part numbers for each case are suffixed "A" and "G" respectively.)

Design Note 30
Issue 1 February 1996

High Speed Turn-off Circuit for PNP Pass Transistors

Cost-Effective replacement of P-Channel MOSFETs

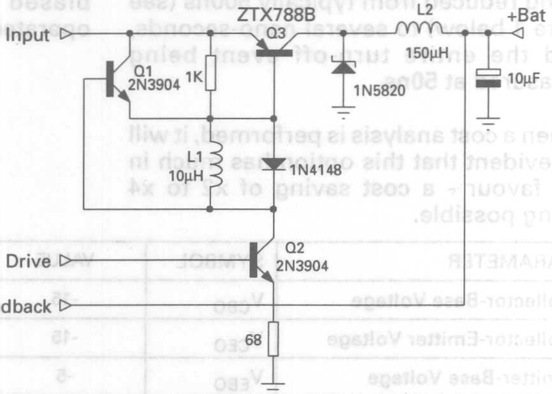


Figure 1
Active Turn-off Circuit for High Current, Low $V_{ce(sat)}$ PNP.

The turn-off circuit shown in Figure 1 can be used with an efficient PNP transistor as a cost effective alternative to P-Channel MOSFETs in step-down (Buck) converters.

It is well appreciated that optimised Bipolar transistor technologies can display lower on-state loss than similar sized (or indeed much larger) MOSFETs, thereby leading to lower cost switch options. [This is essentially due to the pattern of current flow within the structure, and allows relatively small die geometries to display current capabilities far in excess of like packaged MOSFET options.]

However at high switching frequencies, the turn-off times of the Bipolar device, and particularly the storage time portion of the turn-off transition, can lead to significant switching power losses, limiting the product's operation at frequencies much above 70kHz. The MOSFET (being a majority carrier device) does not exhibit storage time effects, and so simpler passive turn-off networks can be employed at low-medium switching frequencies (~100kHz) providing turn-on and turn-off times in tens of nano-seconds. (Note : Though for optimum performance, and especially at higher switching frequencies, attention must be given to the control of gate charge, and the gate drive circuitry must be capable of sourcing/sinking high charge/discharge currents.)

The addition of a few inexpensive components to effect this turn-off circuit allows low loss operation to 150kHz and beyond. Comparison of turn-off times with passive turn-off circuitry, and with this modification show a dramatic improvement. Storage time particularly being reduced from typically 500ns (see Note 1 below) to several nano-seconds, and the entire turn-off event being measured at 50ns.

When a cost analysis is performed, it will be evident that this option has much in it's favour - a cost saving of x2 to x4 being possible.

The ZTX788B is one member of the ZTX788B - 796A Super-β PNP family, which are also available in the SOT223 surface mount package.

Note 1: Measured on a simple PWM IC controlled, 12 to 5V,100kHz converter, biased for 1A output, and the PNP operated with a forced gain of 50.

PARAMETER	SYMBOL	VALUE	UNIT
Collector-Base Voltage	V_{CBO}	-15	V
Collector-Emitter Voltage	V_{CEO}	-15	V
Emitter-Base Voltage	V_{EBO}	-5	V
Peak Pulse Current	I_{CM}	-8	A
Continuous Collector Current	I_C	-3	A
Practical Power Dissipation*	P_{totp}	1.5	W
Power Dissipation at T_{amb}=25°C derate above 25°C	P_{tot}	1 5.7	W mW/°C
Operating and Storage Temperature Range	tj:tstg	-55 to +200	°C

		MIN.	MAX.		CONDITIONS.
Collector-Emitter Saturation Voltage	$V_{CE(sat)}$		-0.15 -0.25 -0.45	V V V	I_C=-0.5A, I_B=-2.5mA* I_C=-1A, I_B=-5mA* I_C=-2A, I_B=-10mA*
Static Forward Current Transfer Ratio	h_{FE}	500 400 300 150	1500		I_C=-10mA, V_{CE}=-2V* I_C=-1A, V_{CE}=-2V* I_C=-2A, V_{CE}=-2V* I_C=-6A, V_{CE}=-2V*

Table 1
Maximum Ratings and Sample Parametric Data for ZTX788B Transistor - (For full details please refer to the datasheet).

ZETEX

Design Note 31
Issue 2 November 1995

High Voltage Generation for Xenon Tube Applications

Introduction

The ignition timing lights in common use range from simple neon to complex units. Neon timing lights have a drawback that due to their low light output, the user is forced to operate them in subdued lighting. This becomes a safety hazard as one tends to hold the unit close to the timing mark and to the "invisible" (or apparently stationary) fan blades.

The ignition lights which use Xenon filled stroboscopic tubes are much better, since their light output is of much higher intensity. The circuit described in this note incorporates such a tube, which has an anode voltage rating of 500 volts maximum and requires a trigger voltage between 2-6kV.

The circuit was designed for a four stroke engine. It will be seen later that the unit can be converted for use as a low power stroboscope with some slight modifications.

The required high voltage for the tube was achieved by using an inverter. The inverter must drive a capacitive load and also withstand the secondary being shorted. Operating the inverter in the flyback mode seemed the best choice, since the energy transfer only takes place when the switching transistor is off, thus effectively isolating it from the load.

Circuit Action

In the circuit diagram shown in Figure 1, the transistor Tr7 is the switching device. The transformer T1, converts the voltages and also transfers the energy. The capacitors C10 and C11 are used as energy storage elements. When the tube is triggered, this stored energy is discharged into the tube - which produces a bright flash of light. The brightness of the flash depends on the size of the storage capacitors and the voltage to which they are charged. The switching of Tr7 is controlled by the Schmitt trigger, formed by Tr3 and Tr4, which senses the 'current' through the primary and secondary. On switch on, the transistor Tr3 will be off and Tr4 will be on. Thus, turning on transistor Tr5 pulls the gate of the MOSFET transistor up to approximately the supply voltage. Whilst the MOSFET is on, the energy is stored in the primary inductance of the transformer. The current in the primary increases as a linear ramp whose slope is inversely proportional to the primary inductance. The resistor R11 senses this

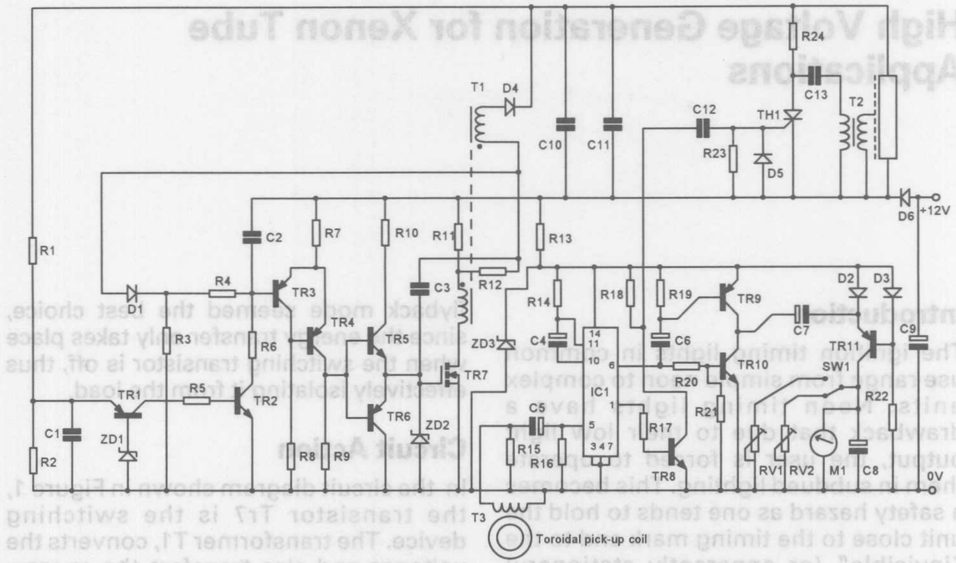


Figure.1
Ignition Timing Light (Parts list in Appendix A)

current and when it reaches a pre-set peak value, sufficient voltage drop is developed across R11 to turn Tr3 on. This in turn switches Tr4 off, which pulls the bases of Tr5 and Tr6 to ground. The gate capacitance of the MOSFET now discharges through Tr6 turning the MOSFET transistor off. When Tr7 turns-off, the primary current immediately ceases and the collapsing magnetic field produces a current ramp of opposite slope in the secondary winding - charging up the output capacitors C10 and C11. The value of this current is the peak value of the primary current divided by the turns-ratio of the transformer T1. This secondary current is also sensed by the same Schmitt trigger circuit. This is achieved by connecting a resistor, R12, in series with the secondary as shown in the diagram. Note that this current also flows through R11.

As the output capacitors C10 and C11 are charged up by the secondary current, the voltage across them gradually increases. Also, as the secondary current ramps down, the voltage drop across R11 and R12 decreases. When the upper threshold voltage of the Schmitt trigger is reached, the transistor Tr3 again turns-off and the next cycle begins. This action continues to dump the energy into the output capacitors until the output voltage reaches the required value. When this has been achieved, the potential divider formed by R1 and R2 and the voltage sensing elements Tr1 and ZD1 inhibit the inverter

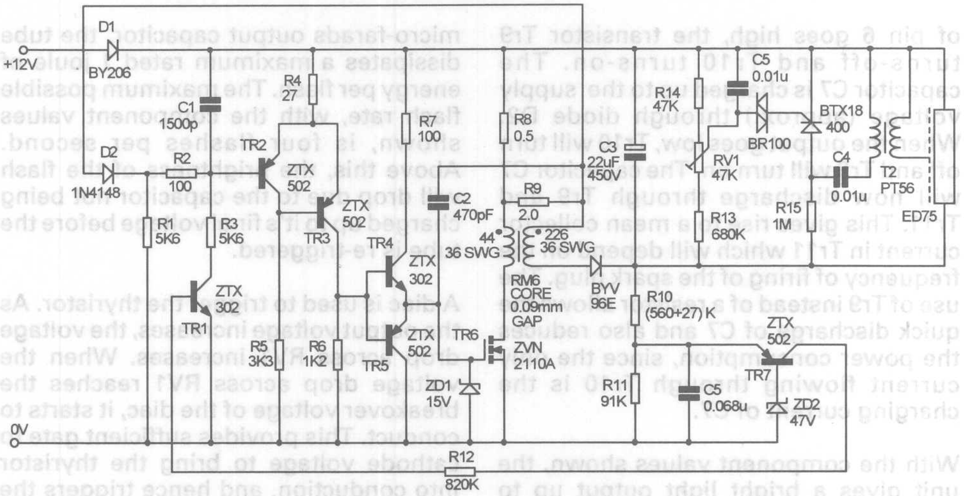


Figure 2
Xenon Beacon

by keeping Tr3 on. When the tube is triggered, the capacitors are discharged and the output voltage drops. The transistor Tr1 turns-off and unlatches the Schmitt trigger and therefore the inverter action resumes.

The use of capacitor C3 enhances the switching of the MOSFET. As Tr7 begins to turn on, the C3-T1 node swings positive by transformer action. This swing is capacitively coupled via C3 to the gate. The regenerative action rapidly switches Tr7 hard on. When Tr7 begins to turn off, the C3-T1 node swings negative. Again the regenerative action rapidly switches the MOSFET transistor off.

The Xenon flash tube is triggered by the firing of the first spark plug in the engine firing order. The transformer T3 is placed over the spark plug and produces a trigger pulse for the monostable every time the spark plug is fired.

A single monostable circuit, a 74121, is used for both the thyristor trigger and a revolution counter. One of the outputs is used to control the transistor Tr8, whose collector is capacitively coupled to the gate of the thyristor. This form of triggering ensures a positive turn-off of the thyristor in each cycle. Hence, the possibility of it remaining in conduction for more than 1 cycle is removed. Prior to the thyristor triggering, the capacitor C13 will be charged to the output voltage. When the thyristor conducts, C13 and the primary inductance of the trigger transformer T2 form an oscillatory circuit. The secondary of T2 produces the required high trigger voltage.

The second output of the monostable is used to drive the revolution counter circuit, whose operation is as follows:

When the spark plug is fired, the output of pin 6 goes high, the transistor Tr9 turns-off and Tr10 turns-on. The capacitor C7 is charged up to the supply voltage (approx.) through diode D2. When the output goes low, Tr10 will turn off and Tr9 will turn on. The capacitor C7 will now discharge through Tr9 and Tr11. This gives rise to a mean collector current in Tr11 which will depend on the frequency of firing of the spark plug. The use of Tr9 instead of a resistor allows the quick discharge of C7 and also reduces the power consumption, since the only current flowing through Tr10 is the charging current of C7.

With the component values shown, the unit gives a bright light output up to about 2500 revolutions per minute, beyond which the light output starts to fall. This is because the output capacitor is not charged to it's final value before it is triggered again. The flash tube dissipates about 4 Watts per flash.

The unit can be easily converted into a low power Stroboscope by simply triggering the monostable with an external square wave oscillator instead of the pick-up coil.

Xenon Beacon

Another possible application for the ZVN2110A MOSFET is a Xenon beacon. Since the beacon flash rate is low, a high value output capacitor can be used. This allows more energy to be dumped into the tube, giving an intense pulse of light. The circuit diagram is shown in Figure 2. The circuit is similar to that described previously.

The output voltage generated in this case is 320 volts and with a 20 micro-farads output capacitor, the tube dissipates a maximum rated 1 joule of energy per flash. The maximum possible flash rate, with the component values shown, is four flashes per second. Above this, the brightness of the flash will drop due to the capacitor not being charged up to it's final voltage before the tube is re-triggered.

A diac is used to trigger the thyristor. As the output voltage increases, the voltage drop across RV1 increases. When the voltage drop across RV1 reaches the breakover voltage of the diac, it starts to conduct. This provides sufficient gate to cathode voltage to bring the thyristor into conduction, and hence triggers the tube. When the tube discharges, the output voltage drops, taking the diac and the thyristor out of the conduction mode and the cycle begins again.

Appendix A
Component Values for Figure 1

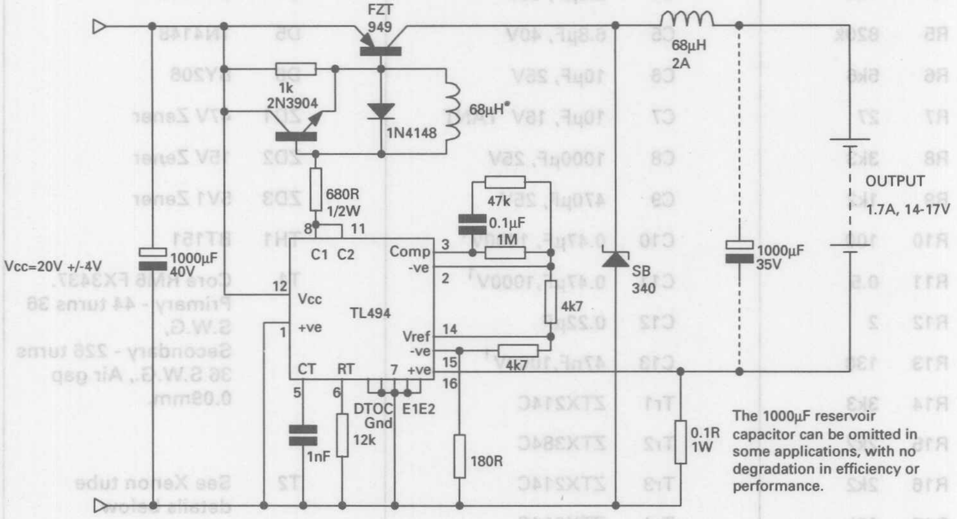
R1	1M	C1	68nF	D1	1N4148
R2	120k	C2	1500pF	D2	ZDX1F
R3	5k6	C3	470pF	D3	1N4000
R4	100	C4	2.2µF, 63V	D4	BYV96E
R5	820k	C5	6.8µF, 40V	D5	1N4148
R6	5k6	C6	10µF, 25V	D6	BY206
R7	27	C7	10µF, 16V TANT	ZD1	47V Zener
R8	3k3	C8	1000µF, 25V	ZD2	15V Zener
R9	1k2	C9	470µF, 25V	ZD3	5V1 Zener
R10	100	C10	0.47µF, 1000V[1]	TH1	BT151
R11	0.5	C11	0.47µF,1000V[1]	T1	Core RM6 FX3437. Primary - 44 turns 36 S.W.G, Secondary - 226 turns 36 S.W.G., Air gap 0.09mm.
R12	2	C12	0.22µF		
R13	130	C13	47nF,1000V[1]		
R14	3k3	Tr1	ZTX214C		
R15	2k2	Tr2	ZTX384C		
R16	2k2	Tr3	ZTX214C	T2	See Xenon tube details below
R17	18k	Tr4	ZTX214C		
R18	4k7	Tr5	ZTX384C		
R19	4k7	Tr6	ZTX214C	T3	Core FX1589, 20 turns, 1mm wire
R20	2k2	Tr7	ZVN2110A		
R21	2k2	Tr8	ZTX300	Xenon Tube ED69 (Integral Reflector and Pulse Transformer T2)	
R22	4k7	Tr9	ZTX214C		
R23	10k	Tr10	ZTX384C	M1	1mA F.S.D., 75Ω
R24	100k	Tr11	ZTX214C		
RV1	100	IC1	74121		
RV2	220				

Note 1: POLYPROPYLENE

ZETEX

Design Note 32
Issue 1 April 1996

A High Efficiency Constant Current Source for Battery Charging Applications



The 1000µF reservoir capacitor can be omitted in some applications, with no degradation in efficiency or performance.

Figure 1
TL494 Constant-Current Source.

This design note describes a 1.7A constant current source for battery charging applications. With an input of 20V, constant current can be generated for loads from 5V to 17V - for charging from 4 to 12 Ni-Cd cell battery packs.

An efficiency of better than 85% is achieved using a very low $V_{CE(sat)}$ bipolar transistor, the FZT949 or the FZT789A, as the switching element.

The circuit uses the Texas Instrument TL494 converter IC (Please refer to Figure 1). A speed up circuit, using a 2N3904 and a 68µH inductor enhances the FZT949/FZT789A switching speed - Collector-to-0V, and drive waveforms are shown in Figure 2.

No heatsink is necessary for the FZT949/ZTX789A due to the high efficiency achieved. However, using the FZT789A will result in a slightly higher

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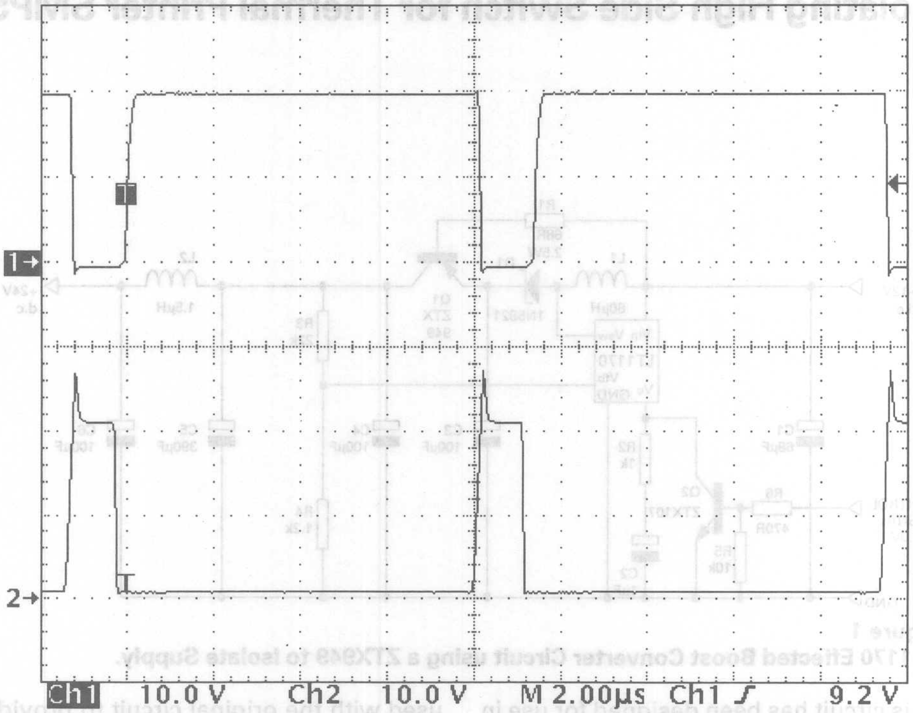


Figure 2
FZT949/FZT789A Switching speed-waveforms.
Upper Trace: Collector-to-0V; Lower Trace: IC Drive.

case temperature than the FZT949 - but still well within the accepted operational specification of the device. If the user prefers to run the case temperature lower, than the FZT949 is the preferred device.

An industry standard switching P-Channel MOSFET (BUZ271 - in TO220 outline) has been substituted for the Zetex device but did not improve on the efficiency - though it does cost considerably more and is only available in a larger package - TO220 compared to SOT223 for the FZT789A/FZT949.

In conclusion, it has been shown in this design note that with the selection of a very low $V_{CE(sat)}$ PNP Zetex bipolar transistor, high efficiencies can be achieved and at a much reduced cost compared to a MOSFET based design.

Isolating High Side Switch for Thermal Printer SMPS

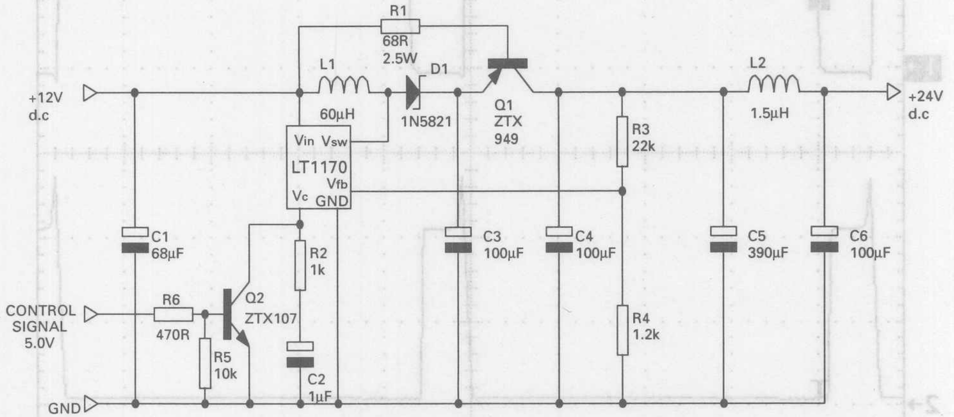


Figure 1
LT1170 Effected Boost Converter Circuit using a ZTX949 to Isolate Supply.

This circuit has been designed for use in thermal printers, to isolate the output of a boost converter from the input supply.

This allows the print head to be powered down when not in use, so that the thermal paper is not discoloured - this can be a problem with this converter topology as even when the converter IC is disabled, a direct path is available through the boost inductor L1 and the Schottky diode D1.

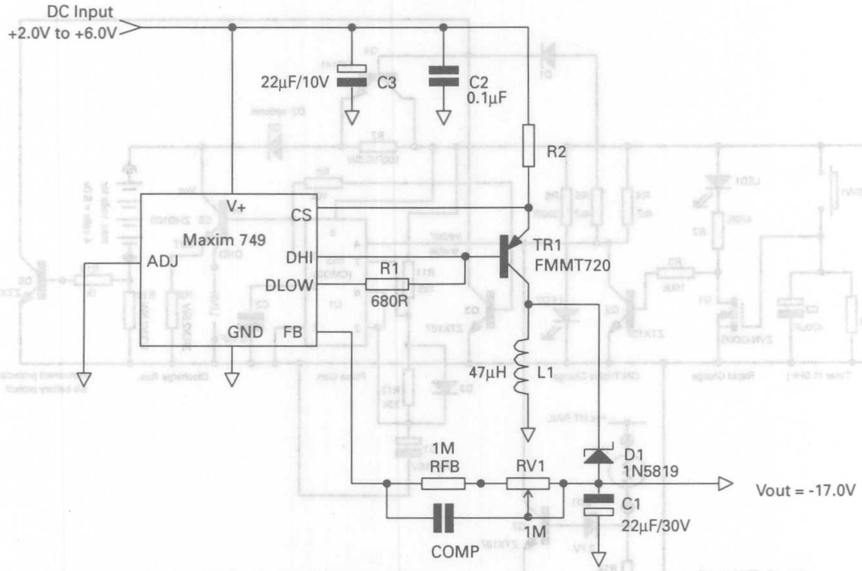
The circuit modification includes a low $V_{CE(sat)}$, high current PNP transistor - the ZTX949, as a high side switch controlled via the ZTX107. These components are used with the original circuit to provide an isolated 24V 2A high efficiency power supply.

The low $V_{CE(sat)}$ of the ZTX949 (rated at a BV_{CEO} of 30V, a continuous collector current of 4.5A and a pulsed current at 20A) enables the device to switch low duty cycle pulse trains up to 7A without a heatsink.

Similar isolating switches can be configured using P-Channel MOSFETs, but due to their lower silicon efficiency these parts do not permit cost and space effective designs.

Design Note 34
Issue 1 April 1996

High Efficiency Converter for LCD Bias Contrast Supplies



High Efficiency Negative Boost Converter.

A high efficiency power supply based on the Maxim MAX749 for generation of the LCD bias contrast voltage is presented in this design note. The circuit requires an input of +2V to +6V and delivers an output of -17V at 50mA.

It is ideally suited for portable equipment and can be powered by two to five 'AA' size Ni-Cd batteries.

Efficiency in the order of 80% is achieved using the FMMT720 - one of the very low $V_{CE(sat)}$ SOT23 PNP transistors available from the Zetex SuperSOT series.

Design Note 35
Issue 1 April 1996

Low Cost Ni-Cd and Ni-MH Fast Charge Circuit

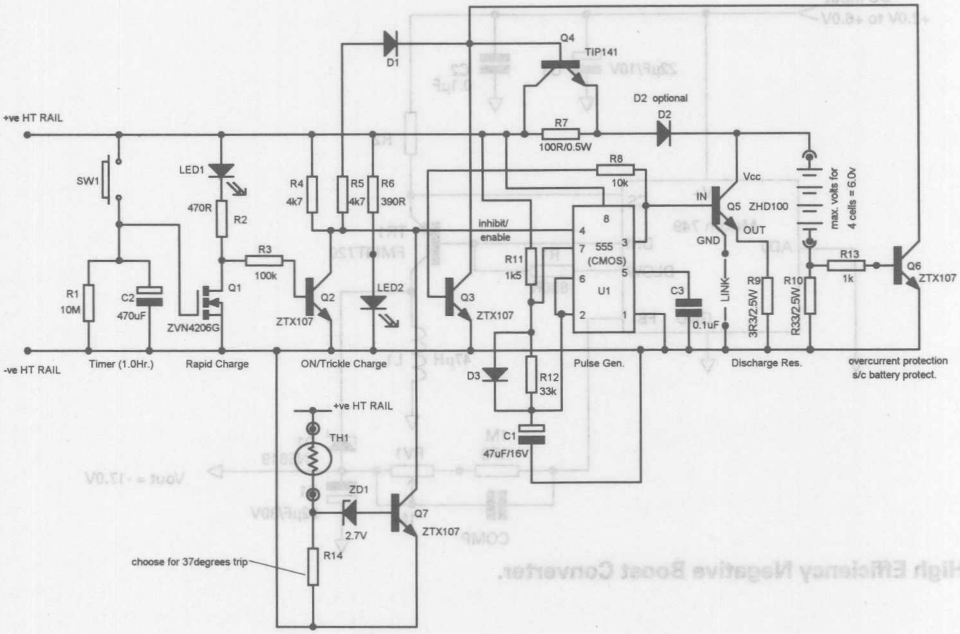


Figure 1
Low Cost Battery Charging Circuit.

Figure 1 shows an economical charger which has been designed for fast charging of Ni-Cd and Ni-MH battery packs, and also has been shown to reduce the 'memory effect' in Ni-Cd cells.

Generally, Ni-Cd packs are trickle charged at 1/10C for about 14 Hrs. With increased understanding of battery chemistry the trend is shifting towards rapid charging at higher rates, e.g. 1C and greater, especially in the professional market. (Figure 2)

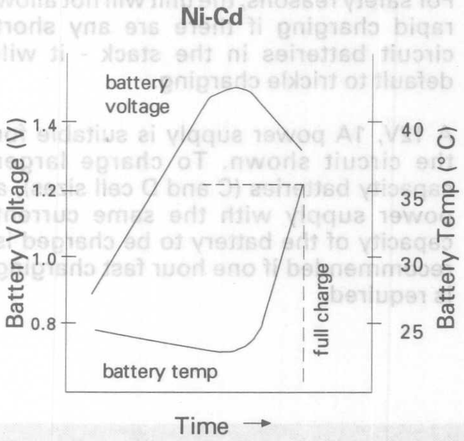


Figure 2
Charging Characteristics of Ni-Cd Batteries.

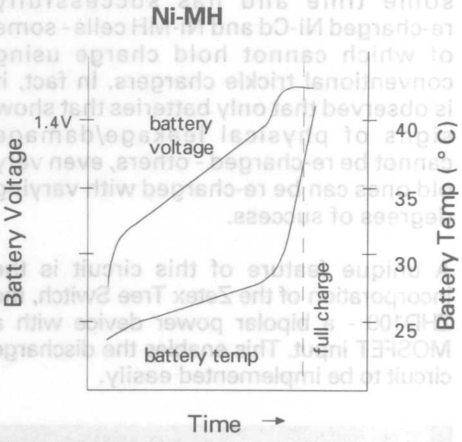


Figure 3
Charging Characteristics of Ni-MH Batteries.

The new generation of 'smart charger' employs an ASIC, often in combination with a microcontroller to optimize battery management. The methods used to detect end of charge are dV/dt inflexion, temperature and time.

In this simple design, it is found that monitoring of temperature to detect end of charge can be as effective as the dV/dt method - which can be a problem for Ni-MH (see Figure 3) as the inflexion point is not well defined. However, as the temperature at end of charge for Ni-MH (38°C) is greater than Ni-Cd (35°C), using the temperature only method will result in a slight undercharge for Ni-MH batteries.

The circuit presented can charge 4 x 'AA' size batteries in series within 1 hour, - less if the batteries are not fully discharged. The circuit supplies a charge (3s), then discharge (10ms) current waveform when rapid charging. This repeated discharging during the charge cycle can reduce (and in some cases remove) the effect of 'memory retention' in Ni-Cds.

At switch on, the charger defaults to trickle charging at 70mA (so can be used as a simple conventional charger). When SW1 is momentarily closed a simple 1.2 hour timer is enabled and the charger goes into rapid charge mode - 1C rate (about 1A). The temperature of the battery rises as it nears full charge and when it reaches 35°C, the unit reverts to trickle charge and stays indefinitely in this maintenance mode - until SW1 is closed again.

The prototype unit was set such that the charge duration is greater than that required to charge Ni-MH from zero depth of discharge. This ensures the batteries will be charged to maximum capacity whatever the initial state of the batteries. The unit has been in use for some time and has successfully re-charged Ni-Cd and Ni-MH cells - some of which cannot hold charge using conventional trickle chargers. In fact, it is observed that only batteries that show signs of physical leakage/damage cannot be re-charged - others, even very old ones can be re-charged with varying degrees of success.

A unique feature of this circuit is the incorporation of the Zetex Tree Switch, the ZHD100 - a bipolar power device with a MOSFET input. This enables the discharge circuit to be implemented easily.

For safety reasons, the unit will not allow rapid charging if there are any short circuit batteries in the stack - it will default to trickle charging.

A 12V, 1A power supply is suitable for the circuit shown. To charge larger capacity batteries (C and D cell sizes), a power supply with the same current capacity of the battery to be charged is recommended if one hour fast charging is required.

ZHD100 DEVICE DESCRIPTION

The ZHD100 device is a monolithic high side switch intended for automotive and industrial applications. Accepting a control signal from standard 5V CMOS/TTL logic families, the device can switch grounded loads of up to 250mA continuous. With a maximum supply voltage rating of 80V, the ZHD100 can withstand a variety of automotive hazards including load dump transients.

FEATURES

* High side switch
* 5V CMOS/TTL compatible input
* 80V max supply voltage
* 250mA max continuous output current
* 5V reverse supply blocking capability
* 4 lead SOT223
* High input impedance
* Low on-voltage drop

APPLICATIONS

* Lamp driver
* Relay driver
* Motor driver
* Solenoid driver
* Diagnostic systems

ZETEX

Design Note 36
Issue 1 May 1996

ZDC833A Dual Variable Capacitance Diode

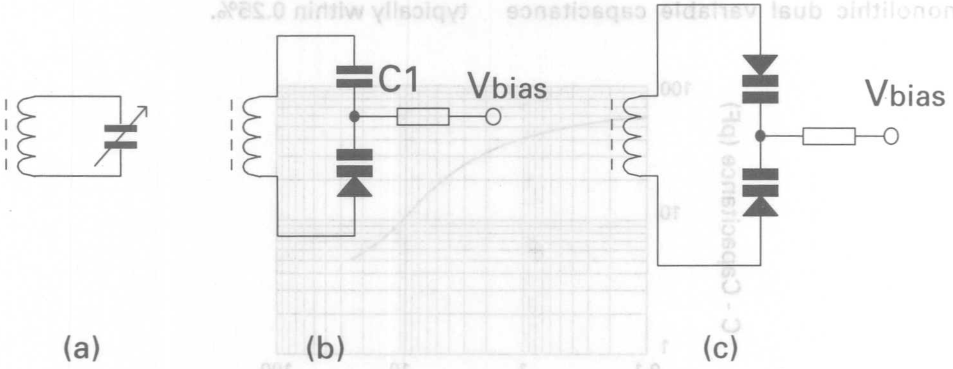


Figure 1.
Basic Tuned Circuits.

Variable capacitance diodes have established a broad applications base from voltage controlled crystal oscillators (VXCO) to television and cellular communications. They are capable of providing many advantages in terms of space and cost savings, and also allow implementation of such features as frequency scanning and stored tuning.

The standard 'LC' parallel circuit is shown in figure 1a). For electronically tuned systems the 'C' element is replaced by a variable capacitance diode as figure 1b) - this design requires a DC blocking capacitor C1, and a voltage bias supply. This method does have some disadvantages, namely:

1. Under high signal conditions, the diode can become forward biased leading to rectification and generation of distortion products.

2. The maximum frequency of operation is limited by the minimum capacitance exhibited by a single diode.

3. The DC blocking capacitor required needs to have a large value which may be undesirable in space limited designs.

Many oscillator circuits are designed for the higher frequency bands where low values of capacitance are required. In these cases, a common option is to use two variable capacitance diodes in series as shown in figure 1c).

This provides a circuit capacitance of a half that of a single diode, prevents forward biasing (as each diode is blocked by its reverse connected partner), and removes the requirement for a DC blocking capacitor.

To allow further cost and space reductions, Zetex have produced a monolithic dual variable capacitance diode available in the industry standard SOT23 package. This part is based on the single diode ZC833A device, and so has been called a ZDC833A - the capacitance versus voltage curve (C-V) has been reproduced in figure 2. As the dual diode is produced monolithically (Eg. on the same piece of silicon), the parametric matching is inherently excellent, typically within 0.25%.

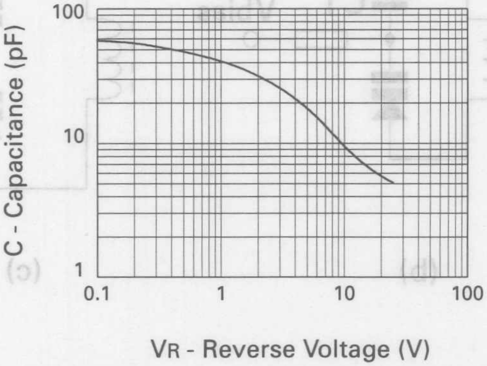


Figure 2.
Capacitance v Reverse Voltage for the ZDC833A.

Appendix

PARAMETER	SYMBOL	MIN.	TYP.	MAX.	UNIT	CONDITIONS.
Reverse Breakdown Voltage	V_{BR}	25			V	$I_R = 10\mu A$
Reverse Leakage Current	I_R			20	nA	$V_R = 20V$
Temperature Coefficient	η			400	ppm/°C	$V_R = 3V$, f=1MHz
Diode Capacitance	C_d	29.7	33	36.3	pF	$V_R = 2V$, f=1MHz
Capacitance Ratio	C_d / C_d	5.0		6.5		$V_R = 2V/20V$, f=1MHz
Figure of Merit	Q	200				$V_R = 3V$, f=50MHz

Design Note 37
Issue 1 June 1996

The ZR2431 Low Reference Voltage Shunt Regulator

Shunt Regulator

The ZR2431 is ideal for providing low voltage stabilised supplies or references. Low output voltage supplies can be difficult to design as the components normally used, such as zener diodes or band gap references, are either too poor in performance or not available at the desired voltage (band-gap devices are usually fixed at 1.2V or 2.5V with no intermediate values). The circuit in Figure 1 shows a 1.5V shunt regulator utilising the ZR2431. The output of this supply can be set to any voltage in the range 1.23V to 20V by adjusting the ratio of R1 and R2, following the relationship:-

$$V_{out} = V_{ref} \frac{(R1+R2)}{R1}$$

Where V_{ref} is the reference voltage of the ZR2431.

Note that a small correction to this formula is required if R1 and R2 are given very high values, since the reference input current of the ZR2431 flows through R2 but not R1 causing a small error. This reference current is only 200nA maximum for the ZR2431 so as long as a current of at least 20µA is passed through R1, setting its maximum value at 60k ohms, the reference current can be ignored. For very low current applications where it is advantageous to set R1 and R2 at a level where the reference current must be considered, the following formula takes this input current into account:-

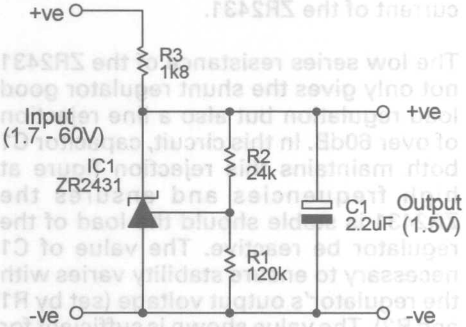


Figure 1
1.5V Shunt Regulator.

$$V_{out} = I_{ref} \times R2 + V_{ref} \frac{(R1+R2)}{R1}$$

Where I_{ref} is the reference input current of the ZR2431.

R3 is selected so as to support the maximum load current at minimum input voltage, yet still maintain adequate operating current for the ZR2431. Since the minimum cathode current of the ZR2431 is only 100µA, it rarely has an effect on the choice of R3.

$$R3 = \frac{V_{in_{(min)}} - V_{out}}{I_{out_{(max)}} + I_{z_{(min)}}}$$

Where $I_{z(min)}$ is the minimum cathode current of the ZR2431.

The low series resistance of the ZR2431 not only gives the shunt regulator good load regulation but also a line rejection of over 60dB. In this circuit, capacitor C1 both maintains this rejection figure at high frequencies and ensures the ZR2431 is stable should the load of the regulator be reactive. The value of C1 necessary to ensure stability varies with the regulator's output voltage (set by R1 and R2). The value shown is sufficient for all voltages but the size of this capacitor can be reduced for higher output voltages e.g. only 47nF is necessary at the highest working output voltage of 20V.

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The ZR2431 Low Voltage Shunt Regulator in SMPS Control Loops

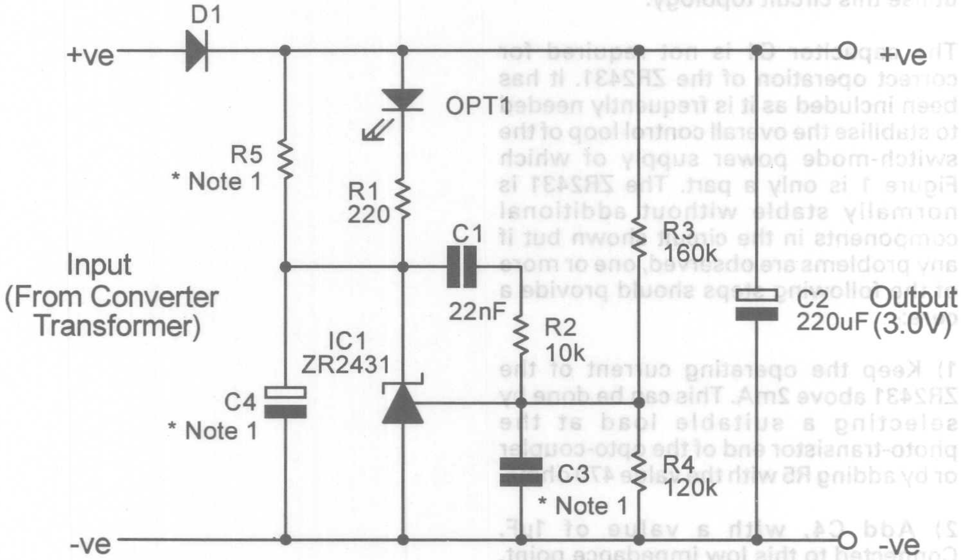

Figure 1
The ZR2431 used within a SMPS Control/Feedback Loop.
Note 1: C3, C4 and R5 are optional components. See text.

Switch-Mode Power Supply Controller

The circuit shown in Figure 1 is commonly used in the control loop of switch-mode power supplies. The output voltage is sensed via R3/R4 by the ZR2431 and the IC controls the current passed through an opto-coupler and hence feeds back output voltage status to the switching regulator. The minimum supply voltage on which this popular circuit can be used is set by the minimum cathode voltage of the regulator IC plus the forward voltage drop of the opto-coupler LED. Previously, the voltage drop of old references possessing a minimum cathode voltage of 2.5V, added to an opto-coupler voltage drop of around 1.2V meant that the circuit was not viable below 4V. Unlike old versions of the 431, the reference voltage of the ZR2431 is 1.23V and its minimum cathode voltage fractionally below this. Consequently, the ZR2431 allows supplies with outputs down to 2.5V to utilise this circuit topology.

The capacitor C1 is not required for correct operation of the ZR2431. It has been included as it is frequently needed to stabilise the overall control loop of the switch-mode power supply of which Figure 1 is only a part. The ZR2431 is normally stable without additional components in the circuit shown but if any problems are observed, one or more of the following steps should provide a cure:-

1) Keep the operating current of the ZR2431 above 2mA. This can be done by selecting a suitable load at the photo-transistor end of the opto-coupler or by adding R5 with the value 470 ohms.

2) Add C4, with a value of 1µF. Connected to this low impedance point, the capacitor should have little if any effect on the control loop of the power supply yet will guarantee stability of the ZR2431.

3) Some supplies use a compensation resistor wired in series with capacitor C1 (in this case R2). If such a resistor has been included, a small capacitor (1nF) located in the position C3 can be an effective stabilising component and may be a preferable alternative to adding the capacitor C4.

Design Note 39
Issue 1 June 1996

ZLDO Series
Quiescent Current vs. Dropout Voltage Trade-Offs

It is a common misconception that low quiescent current is a vital characteristic in a regulator, and dominates the endurance of any battery powered equipment using it. Whilst this is certainly true to a degree, once the factor has been reduced below a reasonable level (dependent on load and battery characteristics), further reduction provides no useful benefits, so improvements should be made in other areas. By comparing the relative gains to be made in minimising quiescent current losses and dropout voltage losses, this note demonstrates how the ultra-low dropout voltage of the Zetex ZLDO regulator series can provide optimum equipment endurance.

Quiescent Current Losses

When considering quiescent current losses, it can be useful to express it as a percentage of the load. To do this, divide the quiescent current of the regulator by (the average load current of the application + regulator Iq), which will give a fraction that represents the loss of operating life caused by quiescent current.

i.e.:-
Efficiency loss = Regulator Iq / (I_{load} + Iq)

The quiescent current of the ZLDO series varies with load current, rising linearly from its zero load value of 550µA by an additional 500µA for the maximum load current of 300mA. With a load current of 60mA, the Iq of a ZLDO500 is around 600µA, losing 0.01 or 1% of equipment operating life to quiescent losses. For a load of 120mA, quiescent current is around 750µA, losing 0.63% operating life and similarly, at 300mA losses would be 0.37%. These percentages represent the maximum gain possible if quiescent current losses could be eliminated completely.

Dropout Voltage Losses

It is an obvious consequence of improving the dropout performance of a regulator that the operating endurance achieved with a given battery source will increase. However, it is fairly difficult to quantify this improvement due to its dependence on cell characteristics and the number of cells used to provide the unregulated supply. The following gives a realistic if not pessimistic example of the advantages of improved regulator dropout performance. Figure 1 shows the discharge curve of a typical Ni-Cd cell operated at a discharge rate of 0.2C.

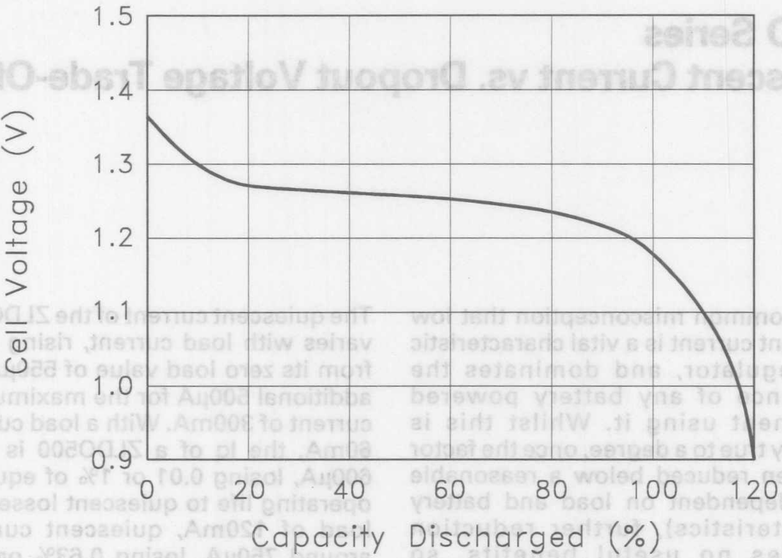

Figure 1
Ni-Cd Cell Voltage During Discharge.

If five of these cells are wired in series to make a battery pack of nominally 6V, the battery could be used as a source for the ZLDO500 5V regulator. This regulator will operate at its maximum specified load current of 300mA with an input voltage down to only 5.1V yet still maintain regulation. This is around 200mV lower than most competing parts and so will allow cells in the battery pack to operate down to a final voltage of 5.1/5 V or 1.02V each, instead of 5.3/5 V or 1.06V as would be permissible for standard LDO regulators. Figure 1 shows that this will increase operating life and this improvement is around 1.5 to 2% for the example discussed.

Quiescent Current/Dropout Voltage Trade-Off

It is immediately apparent that for all but very low loads, the savings that can be made using a low dropout regulator such as the ZLDO series rather than a standard LDO, are far more worthwhile than the gains that could be obtained by reducing or even eliminating quiescent current losses. Also, it should be noted that Ni-Cd cells used in this example have a particularly flat discharge curve with a rapid fall of voltage at end-of-life. For most other battery technologies, the terminal voltage falls more gradually over the cells' life with a much slower fall at end-of-life. With these alternative cell technologies the improvement given by the use of ultra low dropout regulators such as the Zetex ZLDO series will be increased greatly.

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Design Note 40
Issue 1 June 1996

Applications of the ZMR250 Miniature SOT23 Regulator

Adjustable Regulator

There are many occasions when a power supply at a non-standard voltage is required, for which there are no off-the-shelf three terminal regulators available. A common technique of dealing with such requirements, is to use a standard regulator selected with an output voltage lower than needed and to add a potential divider connected to its ground pin to raise the output to the desired level. The power supply circuit shown in Figure1 is an example of this technique.

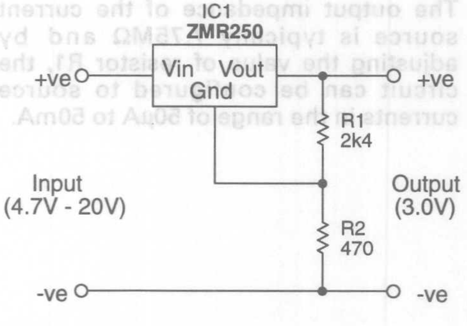

Figure 1
Power Supply with Potential Divider; Designed to Deliver Non-Standard Output.

A serious problem with this circuit, particularly for battery powered equipment, is the current drain of the output potential divider. The resistors selected for this potential divider must be chosen so as to pass around 50 times the quiescent current of the regulator if output accuracy is not to be degraded. Industry standard regulators such as the 78L series usually have quiescent currents in the region of 2mA typical, so to maintain accuracy a divider current of 100mA would be required. However, the quiescent current of the Zetex ZMR250 SOT23 miniature voltage regulator, is only 40µA maximum - even less than that of purpose designed adjustable regulators. Use of the ZMR250 in the above circuit allows the divider string current to be kept down to only 1mA without significantly compromising accuracy, making the power supply viable for even battery powered circuits where efficiency is important.

The output voltage given by the power supply circuit can be calculated using the following formula:-

$$V_{out}=I_q \times R2+V_{reg}\frac{(R1+R2)}{R1}$$

Where Iq is the quiescent current of the ZMR250 (30µA typical) and V_{reg} is the normal output voltage of the ZMR250 (2.5V).

The power supply will operate with inputs from 4.7 to 20V and supply loads up to 50mA (maximum continuous load current may be restricted by package power dissipation).

Current Generator

The low quiescent current rating of the ZMR250 also makes the device very useful for current source applications. Figure 2 shows all that is needed to use the ZMR250 as a current source. This circuit behaves as a two terminal device which passes a current independent of the voltage across it as long as this voltage is in the range of 4.2V to 20V.

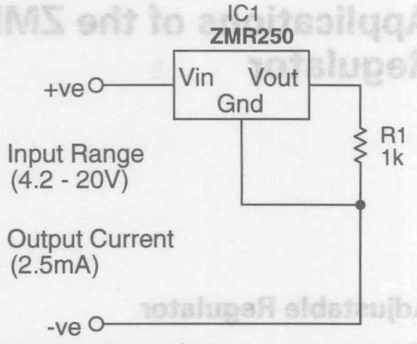

Figure 2
Using the ZMR250 as a Current Source.

The output current is set by resistor R1, following the relationship:-

$$I_{out}=I_q+\frac{V_{reg}}{R1}$$

here I_q is the quiescent current of the ZMR250 (30µA typical) and V_{reg} is the normal output voltage of the ZMR250 (2.5V).

The output impedance of the current source is typically 7.75MΩ and by adjusting the value of resistor R1, the circuit can be configured to source currents in the range of 50µA to 50mA.